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(54) Title: COMMUNICATION SYSTEM AND METHODS OF ESTIMATING CHANNEL IMPULSE RESPONSES THEREIN

(57) Abstract: Multiple Steiner codes are transmitted as bursts (s_{11}, s_{12}, s_{33}, 560, 524) from multiple base stations (182, 184, 186) having one or more transmit elements (174, 176, 178, 180), with successive bursts providing an extended training sequence for use in channel estimation at an addressed unit (172), such as a mobile handset. Accurate channel estimation is possible through the use of Wiener frequency domain MMSE deconvolution (518) combined with frequency domain spatial decoupling matrices, with quasi-orthogonal pseudo-noise sequences (502, 504, 520, 522) allocated to base stations and their antenna elements. The use of Steiner codes to supplement Wiener frequency domain MMSE deconvolution and frequency domain spatial decoupling results in the possibility of allocating only a single training sequence to each base station provided that the training sequence is of sufficient length to encompass all multiple time-translated channel impulse responses (H). Estimates may be refined iteratively by minimising the MS error of demodulated pilot symbols. Estimates may also be refined by removing taps from the impulse response which are insignificant based on a relatively long-term power-delay profile for the channel.

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COMMUNICATION SYSTEM
AND METHODS OF ESTIMATING CHANNEL IMPULSE RESPONSES THEREIN

Description

5   This invention relates, in general, to a communication system and a method of estimating channel impulse response (IR) therein, and is particularly, but not exclusively, applicable to communication environments employing space-time coding. The present invention is also applicable, without imposing limitation, to code division multiple access (CDMA) schemes, orthogonal frequency division

10   multiplexing (OFDM) or the global system for mobile (GSM) communication, and also to systems having base stations configured to transmit on either a time-aligned or unsynchronised basis.

       GSM, CDMA, and OFDM systems and those systems using space-time coding

15   usually require estimation of impulse responses (IR's) between a terminal antenna and several base antennas, especially in instances when the relative signal strengths of the base stations are similar. As will be understood, the physical channel through which propagation occurs can have a severely detrimental effect on the ability to recover data accurately, especially with increasing data rates.

20   Consequently, by knowing the IR, compensation for channel-induced defects can be applied to improve accuracy in data recovery, e.g. through interference cancellation techniques. More especially, with multiple transmit elements at a base station, irrespective of whether there is more than one serving base station, multipath interference results in an inability to resolve individual channels. Indeed, even

25   orthogonally structured data (such as different and time dispersed training sequences) can become cross-correlated (in the face of multipath interference) and hence unusable to resolve individual channels/paths in the context of a deterministic IR approach. Furthermore, it will be appreciated that IR determination is particularly taxing at a handoff point at cell boundaries of, potentially, two or three serving base

30   stations (or Node Bs in the context of CDMA and the Universal Mobile Telecommunications System (UMTS)). At cell-boundary handover, the issue of IR is complicated by the fact that downlink carrier to interference ratios (CIRs) can be zero or negative.

When available, impulse responses can also be used to assist in downlink soft-handover.

The use of matched filtering of standard training sequences to determine channel
5   IR's is considered adequate to drive fingers of a RAKE filter in CDMA applications, such as IS-95 and CDMA2000, given that incoherent soft handoff is also allowed. However, more recent cellular system proposals will stretch matched filtering techniques to their limits and beyond due to the use of higher data throughputs (with more bits per symbol, e.g. higher quadrature amplitude modulation QAM levels)
10   requiring higher energy per bit/noise floor (E_b/N_o) than existing systems, such as IS-95. Such higher data rate cellular systems include the Enhanced Data-rate for GSM Evolution (EDGE) system and multiple-input-multiple-output (MIMO) methods (using multiple transmit and receive antennas) applied to High Speed Downlink Packet Access (HSDPA) standards. MIMO methods potentially may also be applied to
15   UTRA TDD, CDMA2000 and even GSM-EDGE.

In any event, interference cancellation, coherent soft hand-off or space-time coding generally requires a more accurate estimate of channel propagation conditions (than that required with RAKE filtering or matched filtering). In this respect, a long history
20   of interference cancelling base station algorithms (that yield indifferent null-steering performance) lends support to this view.

In terms of proposed 3G (third generation) cellular systems, it is known to use training sequences with a cyclic prefix for pilot tones in orthogonal frequency division
25   multiplexing (OFDM) and CDMA. When a channel has an impulse response of known maximum duration, the use of Steiner codes allows re-use of the same training sequence with a time-offset method to measure simultaneously the channel impulse responses of several users or several antenna with a single fast Fourier transform (FFT) operation. Steiner codes are discussed in the paper "A comparison
30   of uplink channel estimation techniques for MC/JD-CDMA transmission systems" by B. Steiner and R. Valentin, Proceedings IEEE 5th International Symposium on Spread Spectrum Techniques and Applications 1998, Volume: 2, pp. 640-646. Indeed, cyclic Steiner codes are found in the pilot tones for the time division duplex (TDD) component of European UMTS, as presented in the ETSI 3GPP Document
35   TS25.211 "Transport channels and physical channels" v.3.2.0, 1999. In terms of a frame structure for a pilot tone, random data sectors sandwich an observation

window that is preceded by a cyclic prefix of a latter portion of the observation window. Provided that the duration of the cyclic prefix exceeds the channel impulse response, then data in the observation window is a function purely of the pilot sequence (comprised from the combination of the cyclic prefix and data in the

5   observation window) and is not corrupted by spurious unknown data in any way.

The 3G Partnership Project (3GPP) have defined training sequences for TDD (time division duplex) having overall chip lengths of 256 and 512 chips; the overall lengths are derived from 192 chips and 464 chips of basic pseudo-random number

10   sequence codes with, respectively, an additional cyclic prefix of either 64 chips and 114 chips. Each training sequences is designed to allow channel impulse response estimation for differing numbers of users, namely three users in the case of the 256-chip training sequence and eight users in the case of the 512-chip training sequence.

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The training sequences are utilised in the receiving unit to estimate the channel impulse response based on a complex cross-correlation (in real and imaginary phase and amplitude components) between received chips and a local replica of the training sequence. In this regard, it is usual, on a per channel basis, to take a

20   correlation of the channel impulse response (h) with the sequence (s) from the transmitter. More particularly, from a single base station having multiple transmit elements, cyclic offsetting of Steiner codes allows utilisation of a fast Fourier transform (FFT) technique to solve individual channel impulse responses. Steiner cyclic pilot codes can therefore be used in estimating, with a single correlator,

25   channel impulse responses of multiple users that do not mutually interfere. Steiner codes may be Gold codes.

As will be understood, the requirement for cyclic redundancy to mitigate data corruption caused by multipath may be obviated provided that the length of the

30   transmitted training sequence (or pilot tone) is sufficiently long. In generality, cyclic redundancy may be avoided if a training sequence contains a sufficient number of chips, with the overall length of the training sequence determined by the prevailing dispersion conditions associated with the channel. Additionally, for estimating CIR, it is also necessary to consider the physical state of the receiving unit, since the

35   channel for a slow moving or stationary receiving unit (e.g. a handheld device) is ostensibly stable, i.e. constant.

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By way of practical explanation of Steiner code operation from a dual element array of a base station transmitter, it will be appreciated that, in estimating the two channels from the two radiating elements to a single receive antenna, the two
5   radiating elements use a common generic pseudorandom number (PN) training sequence $\{\underline{s}_1, \underline{s}_2\}$ which is end-around shifted (in this case by half a block) to $\{\underline{s}_2, \underline{s}_1\}$ for the second element. The two channel impulse responses are $\{\underline{h}_1\}$, $\{\underline{h}_2\}$ and the receiver antenna sees the superposition of the two convolved sequences:

$$\{y\} = \{s_1, s_2\} \otimes \{h_1\} + \{s_2, s_1\} \otimes \{h_2\}$$

10   where $\otimes$ means a discrete-time convolution operation. When the receiver output is correlated with the PN sequence of the first antenna, the impulse responses of the two distinct channels are recovered in superposed form with a relative time shift of half a block. If the maximum probable duration of the impulse responses is finite and known and less than half the PN code length, the two channel estimates are
15   non-overlapping and orthogonal and can be recovered by slicing an output of a correlator block into two halves. Clearly the process extends easily to estimating M channels by splitting the generic PN sequence into M segments $\{\underline{s}_1, \underline{s}_2,.. \underline{s}_M\}$, where $\{\underline{s}\}$ means a short symbol sequence, and wherein transmissions from the base antenna transmit elements follow the sequences $\{\underline{s}_1, \underline{s}_2,... \underline{s}_M\}$, $\{\underline{s}_M, \underline{s}_1,.... \underline{s}_{M-1}\},...\{\underline{s}_{M-}$
20    $_1, \underline{s}_M,.. \underline{s}_{M-2}\}$.

At the receive antenna, the processing block is correlated with only the first copy, $\{s_1, s_2,.. s_M\}$ of the PN sequence whereupon, at the correlator output, the M different channel impulse responses separate out in time and appear in sequence. Clearly,
25   the cyclic head is different for each downlink path, but this is purely a transmitter burden, whereas the receiver samples the waveform and continues to see the same rotated training sequence. In this way, a common discrete Fourier Transform operation can be used to resolve multiple downlink channel impulse responses.

30   As previously indicated, in order to avoid contamination with unknown data which usually immediately precedes the training sequence, a cyclic copy of the end of each rotated sequence may be prefixed at the start of the pilot block. In other words, when working with, say, 512-point processing blocks and when the duration of the channel impulse response is K samples, the total extent of the transmitted pilot burst

is 512+K samples of which the first K contaminated samples are discarded at the receiver.

5   The maximum number M of channel impulse responses that can be estimated by the Steiner method is:

$$M \leq \frac{N}{K}$$

where N is the length of the training sequence (samples) and K is the maximum likely number of samples in the channel impulse response.

10   Unfortunately, with Steiner codes, the resolution of base stations in different cells is much less attractive. Specifically, whilst it would be possible to continue to expand the Steiner method with ever increasing numbers of built in channels M (e.g. perhaps twice or three times the number needed for one base station), estimating more channels needs longer training sequences and synchronisation of the relevant

15   base stations. Increasing training sequence length, however, is not the main difficulty; a bigger problem is that this Steiner solution would effectively constrain all the bases to use one and the same fundamental PN sequence and this would not allow addressed units to discriminate between bases by different codes when they are searching for hand-off possibilities. Steiner is also inflexible since there is

20   contention over the order in which the bases should cycle their training sequences. This contention will appear at all six interfaces of the base's cell with other cells in a hexagonal layout and so sequence cycling must be set up to be compatible with all adjacent sites that could be affected.

25   As will now be appreciated, communication systems demand the resolution of multiple channels that combine to provide a composite channel impulse response C from observed training sequences S transmitted by several base stations each with multiple antenna elements. In this regard, a time domain channel estimation $\hat{h}_1(t)$ for pilot sequence $s_1(t)$ convolved through channel $h(t)$ that is subject to noise $n(t)$ may

30   be obtained through a time-reversed correlation of the received sequence with a matched filter having the form $s_1^*(-t)$. In the frequency domain, the channel estimation takes the form $\hat{H}_1(\Box_k)$.

        For the purposes of channel estimation, matched filtering of the PN sequence in the

35   receiver can be performed through fast Fourier Transform (FFT) techniques. If the

received data block of N samples is $\{y_1, y_2, .. y_N\}$ (which could be a sequence like $\{\underline{s}_1,$ $\underline{s}_2,.. \underline{s}_M\}$), then FFT operation yields:

$$\{s_0, s_1 ... s_{N-1}\} \overset{FFT}{\Rightarrow} S$$
$$\{y_0, y_1 ... y_{N-1}\} \overset{FFT}{\Rightarrow} Y \quad ,$$

In other words, matched filtering operation in the time domain is equivalent to a point-by-point multiplication in the discrete Fourier domain, namely:

$$\hat{H}_k^{(MF)} = Y_k \cdot S_k^* \quad k = 0...2^n - 1$$
$$\hat{H}_k^{(MF)} \overset{IDFT}{\Rightarrow} \{\hat{h}_0^{(MF)}, \hat{h}_1^{(MF)} ... \hat{h}_{N-1}^{(MF)}\}$$

with two different channel estimates contained in the semi-sequences of $\{\underline{h}\}$, namely:

$$\{\hat{h}\}_1 = \{\hat{h}_0, \hat{h}_1 ... \hat{h}_{N/2-1}\}$$
$$\{\hat{h}\}_2 = \{\hat{h}_{N/2} ... \hat{h}_{N-1}\}$$

Of course, once in the frequency domain, sidelobes present in the frequency spectrum in a channel estimation circuit containing a matched filter may be minimised through the use of a Wiener (least squares) filter technique.

The Wiener least squares filter provides an equalisation technique that utilises a modified inverse filter that controls the white noise response of the filter, i.e. the undesired enhancement of thermal noise from the antenna. The paper "Smart Antennas for Third Generation Mobile Radio Systems" by Martin Haardt (Siemens), Stanford Colloquium on Smart Antennas", July 1999, describes channel equalisation in terms of the Wiener filtering response, and explores how two or more receiving antennas can be included in the description of the received data. A further paper by H Sari et al titled "Transmission Techniques for Digital TV Broadcasting", IEEE communications magazine 33(2) Feb. 1995, discusses channel equalisation in the context of the Wiener filter mechanism. Space-time transmit diversity (STTD) is included in the standards for 3G cellular systems in the European UMTS Terrestrial Radio Access (UTRA) system and similar systems in the American CDMA2000 proposals. Consequently, it is important for these STTD systems to implement efficiently data recovery and equalisation algorithms.

Equalisers and channel whiteners are generally not well conditioned, especially if the channel has zeros or deep minima in its frequency response. The stable

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minimum mean square error (MMSE) Wiener filter solution for the channel is therefore:

$$\hat{H}_k^{(W)} = \frac{Y_k \cdot S_k^*}{\left|S_k\right|^2 + \sigma^2}$$

$$\hat{H}_k^{(W)} \overset{IDFT}{\Rightarrow} \{\hat{h}_0^{(W)} \dots \hat{h}_{N-1}^{(W)}\}$$

where σ^2 is the variance of the thermal noise level in the frequency domain. If the
5   DFT is orthogonal, i.e. a unitary matrix operation, then this σ^2 value is the same as
the time domain noise variance per sample, but most FFT algorithms apply some
form of scaling for which allowance must be made.

Use of the Wiener filter solution results in residual errors caused mainly mainly by
10   thermal noise in the sidelobes (and not error due to imperfections in the code
sequence); this can be contrasted with PN sequence autocorrelation sidelobes
experienced in a matched filter environment.

The paper by J. Blantz et al on the "Performance of a cellular hybrid C/TDMA mobile
15   radio system applying joint detection and coherent receiver antenna diversity", IEEE
J. Selected Areas in Comms. 12(4), May 1994, pp.568-579, describes a multi-user
detection algorithm. A second paper by A. Klein et al on "Zero forcing and minimum
mean square error equalisation for multiuser detection in code-division multiple
access channels", IEEE Trans. Veh. Tech. 45(2), May 1996, pp.276-287, provides
20   further context to the present invention.

In overview, therefore, PN training sequences sent from multiple base stations (or
Node Bs) employing one or more transmit elements are subject to multipath that
results in code cross-correlation and an inability at a receiver to resolve the
25   individual channels and establish the individual channel-specific impulse responses.
More specifically, whilst time alignment of base station transmissions may produce a
summation of signals at the receiver (when employing appropriate windowing on a
chip-by-chip basis with respect to identifiable correlation spikes), the receiver is only
able to detect a composite channel impulse response that is unlikely to reflect
30   accurately any of the actual transmission paths. Moreover, whilst the composite
channel impulse response may be sufficient in the context of soft handover (in IS-
95, for example), the composite channel impulse response is generally insufficient in
third generation systems, including systems offering space-time coding.

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Furthermore, there is a reticence shown by service providers to provide synchronicity between base station transmissions since synchronised transmissions increase infrastructure costs, such as through the necessary provision of an accurate timing reference.

5

It is an object of the invention to overcome disadvantages associated with known methods and devices.

The above object is met by the combinations of features of the main claims, the sub-

10   claims disclose further advantageous embodiments of the invention.

According to a first aspect of the invention there is provided a method of determining channel impulse responses of a plurality of channels to a communication device, the method comprising: performing transform operations on both a replica of a signal

15   sequence s_n and a received training sequence y_n by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins; performing point-by-point operations between corresponding signal

20   sequence frequency bins and training sequence frequency bins; and concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel.

25   Preferably the method further comprises the step of separating training sequence bursts emanating from a single element transmitter by one of a cyclic prefix and a blank (zero) carrier.

Preferably, Steiner codes are transmitted as training sequences, the multiple Steiner

30   codes sent from multiple transmit elements in multiple training bursts. The multiple Steiner codes may be transmitted from multiple elements of a base station transmit antenna.

It is preferred to use a set of matrix operations in the frequency domain to resolve

35   channels to the communication device from multiple transmitting stations, the matrix

operations providing solvable linear equations for the channel impulse response H and the signal sequence S expressible in a matrix-vector form at each frequency bin.

5   A minimum mean square error (MMSE) estimation technique is preferable to solve the linear equations, although other mathematical techniques are possible (as will be appreciated).

The MMSE estimation technique preferably employs a Weiner filtering operation
10   providing:

$$\hat{H} = \left(S^H S + \mathrm{Cov}(B)\right)^{-1} S^H Y$$

where: $Cov(B)$ is a covariance matrix of thermal noise; $S^H Y$ is a matched filter operation arranged to minimize the signal to noise ratio but not to remove mutual interference between channel estimates; and $(S^H S + Cov(B))^{-1}$ is a decoupling matric
15   that removes the coupling between different channels caused by any non-ideal nature of the training sequences.

In a preferred embodiment, the number of bursts sent to the communication device from each transmitting unit in communication contact therewith is calculated as a
20   multiplication of: a number of transmitting elements in a transmit array of a transmitting unit; and a number of transmitting units in communication contact with the communication device.

The training sequence burst between transmitting units may be time-aligned.
25

In a second aspect of the present invention there is provided a method of determining channel impulse responses of channels incident to a communication device, the method comprising: and transmitting multiple quasi-orthogonal pseudo-noise sequences as bursts from multiple base stations each having at least one
30   transmit element, successive bursts providing an extended training sequence for use in channel estimation at the communication device; applying a Wiener frequency domain MMSE deconvolution with frequency domain spatial decoupling matrices to generate channel impulse response estimates for the channels.

Preferably, the quasi-orthogonal pseudo-noise sequences are Steiner codes. It is also preferable to allocate only a single training sequence to each base station, the single training sequence being of sufficient length to encompass all multiple time-translated channel impulse responses.

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In another aspect of the present invention there is provided a communication device operational to receive a plurality of training sequences on a plurality of channels, the communication device comprising: means for performing transform operations on both a replica of a signal sequence s_n and a received training sequence y_n received

10   by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins; means for performing point-by-point operations between corresponding signal sequence frequency bins and training

15   sequence frequency bins; and means for concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel.

20   In another aspect of the present invention there is provided a method of determining channel impulse responses of a plurality of channels established between a plurality of transmitting elements and a communication device in a communication system, the method comprising: substantially simultaneously transmitting different training bursts from each of the plurality of transmitting elements, each burst having a length

25   at least as long as a maximum channel duration in the communication system multiplied by a number corresponding to the plurality of transmitting elements; recovering at the communication device a signal sequence y_n from the different training bursts s_n; and resolving the plurality of channels to recover associated channel impulse responses H for each channel by solving an algebraic matrix

30   operation expressed in matrix-vector form as $Y=SH$, where: S is a matrix of partial training bursts for each channel, each training burst segmented into N pieces in the time domain; Y is a vector of a received signal sequence; and H is a concatenation of different channel impulse response vectors.

35   A further aspect of the present invention provides a computer program product for a processor within a receiver device, the computer program product comprising: code

that performs transform operations on both a replica of a signal sequence s_n and a received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of

5   signal sequence frequency bins and a multiplicity of training sequence frequency bins; code that performs point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency bins; and code that concatenates the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response

10   allowing, in the time domain, generation of the channel impulse response for the channel; wherein the codes reside in a computer readable medium.

According to another aspect of the present invention there is provided a communication device having a receiver coupled, in use, to receive a plurality of

15   channels supporting a signal sequence y_n and training sequence bursts, the communication device having: a signal processing platform to perform transform operations on both a replica of a signal sequence s_n and a received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a

20   channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins; the signal processing platform arranged to perform point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency bins; and the signal processing platform further arranged to concatenate the point-

25   by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel.

30   Preferably multiple Steiner codes are transmitted as the training sequences, the multiple Steiner codes sent to the receiver through multiple channels in multiple training bursts.

Preferably the signal processing platform operates to establish a set of matrix

35   operations in the frequency domain to resolve channels to the communication device from multiple transmitting stations, the matrix operations providing solvable

linear equations for the channel impulse response *H* and the training sequence *S* expressible in a matrix-vector form at each frequency bin.

Preferably the signal processing platform solves the linear equations using a minimum mean square error (MMSE) estimation technique.

It is further preferred that the MMSE estimation technique employs a Weiner filtering operation providing:

$$\hat{H} = \left(S^H S + \mathrm{Cov}(B)\right)^{-1} S^H Y$$

where: *Cov(B)* is a covariance matrix of thermal noise;

$S^H Y$ is a matched filter operation arranged to maximise the signal to noise ratio but not to remove mutual interference between channel estimates; and

$(S^H S + Cov(B))^{-1}$ is a decoupling matrix that removes the coupling between different channels caused by any non-ideal nature of the training sequences.

Yet another aspect of the present invention provides a communication receiver comprising: means for receiving, in use, multiple quasi-orthogonal pseudo-noise sequences as bursts from multiple base stations each having at least one transmit element, successive bursts providing an extended training sequence for use in channel estimation at the communication receiver; and means for applying a Wiener frequency domain MMSE deconvolution with frequency domain spatial decoupling matrices to generate channel impulse response estimates for the channels. Preferably, the quasi-orthogonal pseudo-noise sequences are Steiner codes.

In one aspect of the present invention there is provided a communication device operational to receive a plurality of training sequences on a plurality of channels and a signal sequence y_n, the communication device comprising: a receiver for substantially simultaneously receiving, in use, different training bursts from each of the plurality of channels emanating from a plurality of transmit elements, each burst having a length at least as long as a maximum channel duration multiplied by a number corresponding to the plurality of transmit elements; recovery circuitry for recovering, in use, the signal sequence y_n and the different training bursts s_n; and a processor arranged to resolve the plurality of channels to recover associated

channel impulse responses H for each channel by solving an algebraic matrix operation expressed in matrix-vector form as $Y=SH$, where: S is a matrix of partial training bursts for each channel, each training burst segmented into N pieces in the time domain; Y is a vector of a received signal sequence; and H is a concatenation
5   of different channel impulse response vectors.

According to the present invention, a base station of a communication system may be provided in which base station a transmitter chain is arranged to transmit multiple quasi-orthogonal pseudo-noise sequences as training bursts s_n from at least one
10   transmit element and further arranged to transmit a signal sequence y_n, successive training bursts providing an extended training sequence for use in channel estimation at a communication device of the communication system, the transmitter chain substantially simultaneously transmitting, in use, different training bursts from each of the at least one transmit element, each training burst having a length at
15   least as long as a maximum channel duration in the communication system multiplied by a number corresponding to a plurality of channels to the communication device, the extended training sequence and the signal sequence y_n providing a resolution mechanism to the communication device allowing the communication device to resolving the plurality of channels to recover associated
20   channel impulse responses H for each channel by solving an algebraic matrix operation expressed in matrix-vector form as $Y=SH$, where: S is a matrix of partial training bursts for each channel, each training burst segmented into N pieces in the time domain; Y is a vector of a received signal sequence; and H is a concatenation of different channel impulse response vectors.
25

Preferably, the transmit chain is further arranged to separate training bursts emanating from the at least one transmit element by one of a cyclic prefix and a blank (zero) carrier.

30   It is further preferred that multiple Steiner codes are transmitted as training bursts, the multiple Steiner codes sent from multiple transmit elements in multiple training bursts.

Preferably, the number of training bursts sent to the communication device from
35   each base station in communication contact therewith is calculated as a multiplication of: a number of transmitting elements in a transmit array of a base

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station; and a number of base stations in communication contact with the communication device.

5   Advantageously, the present invention provides a mechanism for resolving a composite channel into a plurality of sub-channels with which specific individual channel impulse responses can be associated with and calculated using a numerically efficient fast transform technique. In other words, irrespective of whether or not there are synchronized base station transmission to an addressed unit from multiple base stations having one or more radiating element, individual channels can

10 be resolved to generate channel impulse responses to aid in data recovery. Beneficially, the present invention is particularly applicable in the estimation of the channel impulse responses between multi-element base stations (in different cells) and the receive antenna array of a subscriber unit located at a 2-way or 3-way handoff point where interference is generally very high.

15

It can be shown that for a dispersive channel having two transmitting base stations, the present invention provides increased accuracy in channel estimation, with smaller errors experienced than in an uncoupled case. Indeed, tests have shown that normalised mobile station errors in channel estimation for the present invention

20 can achieve results of 14.8 decibels (compared with 6.1dB and 7.1dB for techniques solely employing a matched filter and Weiner filter, respectively, in channel impulse response estimation). The present invention, even in conditions of varying base station transmit power, provides an improved accuracy in channel impulse response estimation, which accuracy is reflected in a general measured improvement of about

25 7dB over a range of transmit signal strengths of ±20dB for second base station (relative to a carrier signal from a first base station being set at +3dB above thermal noise at the clip level). The present invention is therefore able to operate in interference and soft handover environments where interference becomes a problem.

30

A solution for estimating several downlink channels from different base antennas to a single terminal antenna by the use of multiple training bursts has been described. The method involves the use of $N \times N$ decoupling matrices in the Fourier domain to solve for N channels and gives superior results compared with attempts to use

35 matched filters designed for added white Gaussian noise (AWGN) channels.

Interference cancellation, space-time coding, and coherent soft hand-off can be enhanced by the use of the present invention.

In another aspect of the invention there is provided a method of estimating a
5   channel impulse response comprising:-

(a)   receiving a signal including a training sequence of predetermined training symbols,

(b)   passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission
10   over a channel,

(c)   demodulating at least one of the training symbols,

(d)   calculating the non quadratic, non linear error metric of the demodulated training symbol against a locally stored record of its known correct value,

15   (e)   adjusting the estimated channel impulse response to substantially minimise the mean square error, and

(f)   feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations

whereby the channel estimates used in the channel equaliser are iteratively
20   refined.

Preferably, the training symbol error metric is calculated as mean square error. The received signal may be a CDMA signal and wherein the training sequence is despread using at least one different spreading factor to recover training sequence
25   bits or sub-bits.

It is further preferred that multiple transmissions are received over a plurality of channels each including a training sequence and wherein the summed mean square error of all training sequences is minimised by adjusting the estimated channel
30   response h for each tap of each channel impulse response in turn according to the equation

$$\theta(h) = (h - h_{OPT})^{H} M (h - h_{OPT}) + c$$

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where $\underline{h}$ is the vector of channel impulse responses for all channels, h_{OPT} is the optimum channel impulse response, M is a matrix representing cross-coupling between channel components and c is the minimum error.

5   It is yet further preferred that multiple transmissions are received over a plurality of channels each including a training sequence and wherein the summed mean square error of all training sequences is minimised by adjusting the estimated channel response h using a technique selected form gradient descent methods, search methods, and direct solution algorithms.

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Preferably, $\theta\ (h)$ is minimised by calculating $\theta(h)$ and $\theta(h\pm\delta)$ where $\pm\delta$ is a relatively small symmetric perturbation about h, fitting a curve to the three points generated thereby and finding the minimum of the fitted curve in terms of h.

15   The channel equaliser may be a Wiener (MMSE) equaliser.

It is further preferred that a first estimate of the channel response is generated by:- performing transform operations on both a replica of the transmitted training sequnce s_n and a received training sequence y_n received by the communication
20   device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins; performing point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency
25   bins; and concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the first estimate of the channel impulse response for the channel.

30   According to another preferable feature, the estimated impulse response is further refined by obtaining a long-term averaged power delay profile for the channel, setting a predetermined variance threshold, estimating a channel impulse response having a plurality of taps, and removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a
35   variance below the predetermined variance threshold.

The predetermined variance threshold may be a power threshold or an accuracy threshold.

According to another aspect of the invention, there is provided a method of
5   estimating a channel impulse response comprising:-

    (a)   obtaining a long-term averaged power delay profile for the channel,

    (b)   setting a predetermined variance threshold,

    (c)   estimating a channel impulse response having a plurality of taps, and

    (d)   removing taps from the channel impulse response estimate
10   equivalent to those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

The predetermined variance threshold is preferably dependent on the level of interference in the channel.
15

There are a number of methods for obtaining the long term power delay profile, including, but not limited to obtaining the long term power delay profile based on knowledge of the likely maximum time delay in any given channel or by using a simple estimate of the channel impulse response or from historical channel impulse
20   responses for the channel or by deriving a channel impulse response from an autocorrelation function of the received signal.

It is further preferred that the channel impulse response estimate is combined with its accuracy as predicted point-by-point from the combined power delay profile and
25   interference level, whereby a modified short term channel estimate is produced in which points likely to have poor accuracy are deweighted and those likely to have good accuracy are retained such that final equalise performance is optimised. Preferably, the short term channel estimate is produced using a Bayesian method.

30   Preferably, the channel impulse response $\hat{H}$ is estimated according to the equation

$$\hat{H}_{MMSE} = Diag\left(\frac{\lambda + \rho_k}{\lambda \rho_k}\right)^{-1} Diag\left(\frac{\lambda + \sigma^2}{\lambda \sigma^2}\right)\hat{H}_1$$
$$= Diag\left(\frac{\rho_k}{\sigma^2} \cdot \frac{\lambda + \sigma^2}{\lambda + \rho_k}\right)\hat{H}_1$$

18

where ρ_k is a long-term standard deviation of each channel estimate tap which provides a long-term power delay profile for the channel, λ is the estimation standard deviation of the least squares estimator, and σ is the standard deviation of

5   thermal noise in the channel.

According to a further aspect of the invention, there is provided a method of estimating a channel impulse response comprising:-

(a)   taking long-term measurements to build a database of where in time,
10   multipath interference is located in the channel to generate an
                incoherent power delay profile for the channel, and

(b)   modelling a relatively short term channel impulse response by
        allocating variable complex amplitudes and phases to the taps at the
        locations in the short term model which have the multipath power in
15   the incoherent power delay profile which is above a predetermined
        threshold.

Preferably, the channel impulse response having a plurality of taps is estimated by performing transform operations on both a replica of a signal sequence s_n and a

20   received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins;

25   performing point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency bins; and

        concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response

30   for the channel.

It is further preferred that the channel impulse response estimates are refined by receiving a signal including a training sequence of predetermined training symbols, passing the received signal through the channel equaliser to substantially remove

35   distortion of the signal caused by transmission over a channel, demodulating at least one of the training symbols, calculating the mean square error of the demodulated

training symbol against a locally stored record of its known correct value, adjusting the estimated channel impulse response to substantially minimise the mean square error, and feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations whereby the

5   channel estimates used in the channel equaliser are iteratively refined.

According to a further aspect of the invention, there is provided a computer program product which when executed by a computer causes the computer to carry out the steps of:-

10 (a)   receiving a signal including a training sequence of predetermined training symbols,

(b)   passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission over a channel,

(c)   demodulating at least one of the training symbols,

15 (d)   calculating a non quadratic, non linear error metric of the demodulated training symbol against a locally stored record of its known correct value,

(e)   adjusting the estimated channel impulse response to substantially minimise the mean square error, and

(f)   feeding the adjusted estimated channel impulse response back to the

20   channel equaliser for use in subsequent equalisation operations
whereby the channel estimates used in the channel equaliser are iteratively refined.

It is further preferred that the training symbol error metric is calculated as a mean square error and that the product further includes discarding channel estimate taps

25 which have poor accuracy as predicted from the channel power delay profile.

According to another aspect of the invention, there is provided a computer program product which when executed by a computer causes the computer to carry out the steps of:-

30 (a)   obtaining a long-term averaged power delay profile for the channel,

(b)   setting a predetermined variance threshold,

(c)   estimating a channel impulse response having a plurality of taps, and

(d)   removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a variance below the

35 predetermined variance threshold.

20

Preferably the product includes iterative adjustment of the channel estimates to minimise the mean square error on at least one training symbol.

5   In another aspect of the invention, there is provided a communications receiver arranged to receive signals from a plurality of sources and arranged to estimate a channel impulse response for the channel between the receiver and each source by:-

(a)   receiving a signal including a training sequence of predetermined training symbols,

10   (b)   passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission over a channel,

(c)   demodulating at least one of the training symbols,

(d)   calculating a non quadratic, non linear error metric of the demodulated training symbol against a locally stored record of its known correct value,

15   (e)   adjusting the estimated channel impulse response to substantially minimise the mean square error, and

(f)   feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations

whereby the channel estimates used in the channel equaliser are iteratively refined.

20

Preferably the receiver further includes discarding channel estimate taps which have poor accuracy as predicted from the channel power delay profile and the training symbol error metric is calculated as a mean square error.

25   According to a further aspect of the invention, there is provided a communications receiver arranged to receive signals from a plurality of sources and arranged to estimate a channel impulse response for the channel between the receiver and each source by:-

(a)   obtaining a long-term averaged power delay profile for the channel,

30   (b)   setting a predetermined variance threshold,

(c)   estimating a channel impulse response having a plurality of taps, and

(d)   removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

35

Preferably the receiver includes iterative adjustment of the channel estimates to minimise the mean square error on at least one training symbol.

In further aspects, the invention provides methods, apparatus or computer program
5   products which produce a channel impulse response estimate by receiving a signal including a training sequence of predetermined training symbols, passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission over a channel, demodulating at least one of the training symbols, calculating the mean square error of the demodulated training
10   symbol against a locally stored record of its known correct value, adjusting the estimated channel impulse response to substantially minimise the mean square error, and feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations whereby the channel estimates used in the channel equaliser are iteratively refined.
15

In yet further aspects, the invention provides methods, apparatus or computer program products which produce a channel impulse response estimate by obtaining a long-term averaged power delay profile for the channel, setting a predetermined variance threshold, estimating a channel impulse response having a plurality of taps,
20   and removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

The summary of the invention does not necessarily disclose all the features
25   essential for defining the invention; the invention may reside in a sub-combination of the disclosed features.

At least one exemplary embodiment of the present invention will now be described with reference to the following drawings, in which:
30   FIG. 1 is a schematic block diagram of a basic transmitter and receiver architecture 10 in which processing circuitry may be adapted to support the underlying principles of the present invention;
         FIG. 2 is a diagrammatic illustration of a typical cellular communication system;

FIG. 3 shows a mechanism and schematic block diagram for channel impulse response estimation that may be adapted to support the principles of the preferred embodiments of the present invention;

FIG. 4 is a transmission regime for a training sequence according to a preferred embodiment of the present invention;

FIG. 5 a schematic block diagram of the channel estimation process and architecture according to a preferred embodiment of the present invention;

FIG. 6 is a Steiner code variant of FIG. 4;

FIG. 7 is a schematic diagram of a receiver receiving six training sequences form three base stations;

FIG. 8 is a plot of the Power delay Profile for the ITU Veh-B channel model;

FIG. 9 is a schematic diagram of a diagonalised circulant matrix;

FIG. 10 is a plot showing the significance weighting of a channel tap estimate;

FIG. 11 is a schematic block diagram showing iteration of a Fourier domain solution to minimise pilot estimate MSE;

FIG. 12 is a plot of the functional relation between a channel tap value and MS filter error;

FIG. 13 is a plot of variation in MS error via a single channel coefficient;

FIG. 14 is a plot showing iterative improvement of a 2:2 MIMO channel equaliser in accordance with the invention; and

FIG. 15 is a schematic block diagram of a Fourier domain channel equaliser optimisation.

Detailed Description

The following description is of preferred embodiments by way of example only and without limitation to the combination of features necessary for carrying the invention into effect.

Before discussing the preferred embodiment of the present invention, it should be noted that consideration has been given to employing Steiner codes in combination with Wiener MMSE estimation in the Fourier domain to obtain a low cost multi-user channel estimation method that is highly efficient for simultaneously estimating a number of channels on the downlink for CDMA systems. Steiner codes may also be used in OFDM systems though, in this case, it may be preferred to use the intrinsic

OFDM symbol for training sequences. Given only prior knowledge of the maximum extent of the channel impulse response, the same basic pilot sequence can be time shifted and re-used (with properly determined cyclic headers) to function for a number of different downlink channels, and only one FFT correlation process is

5   needed to solve for all the channels simultaneously. The technique is ideal for wideband fat pipe systems where the same user signal is radiated from different antennas of a transmit diversity antenna or a space-time coded antenna system. Channel estimation by this method is accurate enough to allow adequate interference cancellation of intercell interference, which is usually quite difficult to

10   achieve in CDMA systems due to low grade estimates of the interference parameters.

Referring now to FIG. 1, there is shown a basic transmitter and receiver architecture 10 in which processing logic or circuitry may be adapted to support the underlying

15   principles of the present invention. Transmit and receive chains 100, 101 may be combined in a transceiver arrangement of a station, such as a base station/Node B or a mobile device of a cellular radio system or the like. Individual transmit and receive chains could, of course, be implemented in isolation, whereby unidirectional (as opposed to bi-directional) communication is supported across, for example, an

20   air-interface, such as a CDMA or 3G radio environment (although the present invention is more widely applicable to multi-point transmission systems).

For the sake of explanation only, the following description of the preferred embodiment will be based on a wideband CDMA system and so references to

25   "CDMA" should be construed broadly and without specific limitation.

In transmit chain 100, signal source 102 generates signals 104 which are encoded by encoder 106, such as a half rate turbo coder. Encoded signals 107 are modulated and subjected to CDMA spreading by modulator/spreader 108 and then

30   filtered by pulse shaper digital filter 110. Following filtering, a digital to analogue (D/A) converter is arranged to produce analogue signals 114 which are generally filtered in a second filter 116 to limit any aliasing effect at transmission. A mixer 118 up-converts baseband signals 120 to transmission frequencies prior to amplification in power amplifier 122 and transmission from antenna 124. Of course, as will be

35   appreciated, some of the functional blocks in the transmit chain can be realised as discrete elements or collocated into a single chip or processing block.

Transmitted signals 126 are received by receive antenna 128 of an addressed
station. Received signals 130 are communicated to a down-converting mixer 132 via
an amplifier 134, which down-converting mixer typically produces an intermediate

5   frequency signal for initial signal processing. A filter 134 isolates a desired frequency
spectrum 136 for signal processing, which frequency spectrum is converted to a
digital domain representation 137 by analogue to digital (A-D) converter 138. The
digital domain representation 137 is digitally filtered in digital channel pulse shaper
filter 140 and then applied in series to a combined demodulator/equaliser slicer and

10   RAKE despreader 142 and then a channel decoder 144. Recovered information 145
is ultimately received by terminal receiving equipment (TRE) 146. It will be
understood that channel filtering generally distorts the channel impulse response
and the channel filter becomes part of the unknown channel.

15   The transmit and receive chains 100, 101 are typically micro-processor (□P)
controlled operations. By way of illustration, a controller 150 and associated memory
152 (which may be a combination of random access (RAM) and read only (ROM)) is
shown as being common to both the transmit and receive chains 100, 101. The
memory 152 stores control algorithms, such as equalisation and modulation

20   algorithms, and also operands and incident data is subject to signal processing
requirements. Moreover, common circuitry between the transmit and receive chains
can be shared to reduce component count, subject to the inclusion of suitable
switching and regulating circuitry. Additionally, as will be understood, at least some
of the functions within one or both of the transmit and receive chains 100, 101 can

25   be implemented either entirely or partially as code, e.g. equalisation and
demodulation. It is these code blocks, application specific integrated circuits or
memory modules that can be adapted or up-graded to support the present invention.

Wideband CDMA-specific components can be replaced or substituted to support an
30   arbitrarily selected communication protocol, e.g. an FDD environment employing
TDM techniques or a multicarrier environment.

FIG. 2 is a diagrammatic illustration of a typical cellular communication system 170
showing a subscriber unit 172 (preferably having multiple receive elements 174-180)
35   at cell boundaries between three potentially serving base stations 182-186. For an
equidistant displacement of the subscriber unit from the base stations, one could

25

expect that transmit signal strengths will be roughly equal, subject to the respective paths and the associated physical environments. The general purpose of FIG. 2 is to illustrate a physical environment in which the present invention finds exemplary application. For purposes of clarity, the cellular communication system 170 is shown

5   with three cells 188-192 and a single subscriber unit 172, but clearly cellular systems may include many cells in multiple layers, with each cell servicing a multiplicity of subscriber units. At the hand-off point midway between two or three base stations, it is necessary to distinguish the channel impulse responses for a doubled or tripled number of channels from multiple base antenna elements. If the

10   CIRs are not estimated jointly, the carrier to interference (C:I) ratio is very low within the training sequences and the channel estimates will be correspondingly poor. Given that each base station always has several TX element channels to estimate it is reasonable to use the Steiner method in the intracell mode, allocating one PN sequence per base station with cyclic rotation to distinguish between the bases'

15   antenna elements, as described previously.

Incoming signals 126 to the subscriber unit 172 are received from the antenna 128 at a transmission rate of $^1/T_{ch}$ of the system. By way of practical implementation, an entire slot of a frame is sampled (A/D converted) at the Nyquist rate and read into

20   memory 152 (of FIG. 1) prior to commencement of information recovery. The incoming signals 126 can be received from any digital source of signals and are not restricted to wireless communications, with the base stations utilising single or multiple radiating elements in each transmit antenna array. Signal processing of incoming signals 126 for optimised data recovery is clearly subject to knowledge of

25   the channel impulse response (i.e. the "channel spectrum"), with this determined from training sequences sent from the base stations according to a predetermined regime, e.g. once per slot, once per frame or on any other appropriate basis. An exemplary mechanism for data recovery is described in the aforementioned European and US patent applications (e.g. EP 01300520.2).

30

As regards the channel spectrum, reference is made to FIG. 3 that shows a mechanism for channel impulse response estimation that is adapted to support the principles of the preferred embodiment of the present invention. A training sequence 250, preferably located in and extractable from each slot of a frame, falls incident on

35   receive antenna 128, which training sequence is demodulated in the receive chain (as will be readily appreciated) to recover the random sequence y_k, e.g. a generated

pseudo-random number sequence. The form and structure of the training sequence will be described in detail later, since the structure is associated with an ability to resolve individual channels for channel impulse response estimation according to the present invention.

5

In overview and without initially discussing the preferred embodiment, the recovered random sequence y_k is subjected to a DFT function 252 to produce a channel sequence spectrum Y_k 254. Within the memory 152 associated with the receive chain there is stored a replica of the original training sequence s_k 260, which original

10   training sequence s_k 260 has a corresponding known sequence spectrum S_k 262 obtained by subjecting the original training sequence s_k 260 to an FFT function. The known sequence spectrum S_k 262 may simply be stored in the memory since it is this sequence spectrum S_k 262 that is used to assess the channel impulse response.

15

The DFT function 252 may be performed by a specific digital signal processor (DSP) or the processor 150 of FIG. 1. The channel spectrum therefore assumes the form of an x-point, e.g. 512-point, DFT sample stored temporarily in, for example, RAM.

20   It will be noted that the frequency domain sequence spectrum S_k 262 is essentially flat with slight perturbations about a nominally constant magnitude.

The sequence spectrum S_k 262 is compared with the channel sequence spectrum Y_k 254 in a cross-correlation function 266, such as realised by a Wiener filter

25   frequency response, namely

$$W_k = \frac{H_k^*}{\left|H_k\right|^2 + \sigma^2}$$, to produce an impulse response spectrum H_k 268 that can

be stored in memory for the duration of at least the related slot and longer if desired. Clearly, by storing the impulse response spectrum H_k 268 for an extended period, processing overhead is reduced at the expense of accuracy in any correction of the

30   variations in the channel due to Doppler. As will be understood, time alignment of the received and transmitted training sequences is required prior to cross-correlation.

27

Optionally, the impulse response spectrum H_k 268 may be further refined and so it is subjected to an IFFT function 270 to generate an estimated time domain representation h_k thereof having an arbitrary number of taps/channels that can be used as a matched filter, if desired. In other words, each tap is representative of a
5   weight for the channel impulse response. Within the time domain, trimming of the number of taps reduces the channel impulse response to manageable processing levels, with the reduction reflecting the greatest expected channel impulse response (CIR) duration (presently understood to be seventy-one chips in length in 3GPP, although this number is an arbitrarily selected cut-off). From the reduced number of
10   taps in the channel impulse response, most values will approximate to zero and so a control processor 272, based on a level or number threshold, selects only those N significant taps that most noticeably reflect the channel to produce a best estimate channel impulse response h'_k 274. Selection of the N most significant taps may simply force lesser taps to be forced to zero.   This is explained in more detail below.
15

By limiting the channel impulse response to the most significant taps, processing is reduced and a final error in the channel limited. In the latter respect, it will be understood that all tap values will have an associated error from thermal noise, channel interference and other adverse channel and processing artefacts, so the
20   compounding of errors is mitigated.

A complete understanding of the principles of the present invention will now be gleaned from the following description of specific signal processing interactions and transmission techniques.
25

Concerning transmission of a training sequence according to a preferred embodiment of the present invention, reference is made to FIG. 4. For the sake of explanation, the structuring of base station transmissions (in the timing frame of FIG. 4) is assumed to reflect a three site system in which each of three base stations
30   includes a single transmit element. Of course, the basic principle employed in the specific embodiment of FIG. 4 can be developed for multi-element transmit arrays from two or more base stations and/or multiple receive antennas at a receiver.   FIG. 4 is representative of successive training sequence transmission bursts from each of three base stations (Base TX1, Base TX2 and Base TX3). In generality, the number
35   of training sequences transmitted from all transmit elements follow a squared law, i.e. for three base stations each having single radiating elements then there would

be three successive bursts (t_1-t_3) separated, preferably, by guard periods (g_1 and g_2) or cyclic prefixes. Bursts from the various base stations are preferably time aligned at transmission, although a windowing function at the receiver (that steps through a chip-wise time correlation) may be employed to time align the transmitted

5   sequences s_{nk}. Since channel delay is generally no greater than about 20□s, then the receiver need only step through a chip count commensurate with the maximum channel delay. For example, at an operating frequency (chipping rate) of one megahertz (1MHz), 20□s correspond to twenty chips.

10   During each burst (t_1-t_3), each base station transmits a finite (predetermined) length of training sequence s, which training sequence may vary from 1-chip or more. Typically, each burst will be tens of chips in length and preferably at least about sixty chips in length and most preferably in excess of about one hundred chips. Clearly, the length of the training sequence is restricted to manageable levels to avoid

15   inefficient use of communication resource for more important data transmissions. The nomenclature adopted in FIG. 4 with respect to the bursts follows the convention that the first subscript (n) represents an identity of the transmitting base station whereas the second subscript (k) represents the sequence number. In the context of Base TX1, FIG. 4 is illustrative of three sequences s_{11}, s_{12} and s_{13} during

20   bursts t_1, t_2 and t_3, respectively. A channel sequence spectrum Y_k is formed (for each burst (t_1-t_3)) from the summation of the individual sequences s_{11}, s_{21}, s_{31} from the base stations transmitting in that burst.

For channel estimation, it will be appreciated that the number of samples required is

25   dependent upon bandwidth, e.g. with a symbol rate of 284kbps (kilo bits per second), GSM systems require about five samples of the channel.

It is well known that DFT's have a wrap-round effect that is equivalent to the data being cyclic. Consequently, in (contiguous) observation windows each encoded with

30   the same PN sequence, DFT's can be used to perform a discrete convolution operation. Circulant matrices are equivalent to these DFT operations, which circulant matrices can be defined as follows. In a system in which the training sequence is $S = \{s_0, s_2, s_2, s_3, s_4\}$, the cyclic extension is the channel impulse response is $H = \{h_0, h_1, h_2, h_3, h_4\}$ and the observation window arranged to hold data

35   $Y = \{y_0, y_1, y_2, y_3, y_4\}$, then Y is the cyclic convolution of S with H:

$$y_k = \sum_{n=0}^{N} s_k h_{k-n\,\text{MOD}\,N}$$

with N=4. If the associated DFT's are the sequence spectrum S_k, the channel impulse response H and the channel sequence spectrum Y_k is then:

$$Y_k = S_k \otimes H_k$$

5   where $\otimes$ is a point-by point-product.

In developing a solution to the problem of channel estimation in the presence of interference or multi-user channel estimation (in either non-dispersive or highly dispersive channels), the inventor has appreciated that the fundamental problems

10   associated with channel isolation/resolution can be expressed algebraically in similar terms to multi-user detection algorithms, as referenced above by J. Blantz et al and A. Klein et al.

In the Steiner code method described above, to estimate M channels the length of

15   the PN code must be at least M times the maximum channel dispersion. In an N base station situation, each base station with M elements, while the Steiner method can continue to be used, the PN code length must be at least MN times the channel impulse response duration and each element of each base station would transmit an end-around rotated version of the code with a different unique offset. In the multiple

20   base station situation, the differing round-trip propagation times to the subscriber unit must also be taken into account in the channel model so codes becomes increasingly extended. A more flexible option for allowing expansion of the number of serving base stations is to use an alternating sequence of different Steiner codes, each of which on its own is only just long enough for the number of elements in a

25   single base station to be resolved.

To appreciate the present invention, it is preferable to consider a communication system having N active/serving base stations for a call, with each base station having a single transmitting element. If the N base stations each with a single

30   element are received by a single element antenna of a subscriber unit, such as cellphone or terminal, then if the N bases simultaneously send N strings of N different pseudorandom bursts and such that there are N^2 different sequences altogether, then the individual channel impulse responses can be resolved at the

terminal. In practice, the present invention is particularly interested in a three-base station situation which, statistically, occurs most frequently.

The base stations are preferably synchronised to a time accuracy much less than

5   the duration of any cyclic prefix (which is typically of the order of about 10□s to 20□s for urban microwave propagation). The mth received burst is now the convolution:

$$y_t^{(m)} = \sum_{k=0}^{K} \sum_{n=1}^{N} s_{n,t-k}^{(m)} h_{n,k} + b_t^{(m)} \qquad t = 1,...T, m = 1...M$$

where b is the thermal noise component and K is the maximum dispersion in chips.

10   After transformation into the DFT (frequency) domain, the equivalent, numerically efficient function can be expressed in the point-by-point product form:

$$Y^{(m)} = \sum_{n=1}^{N} S_n^{(m)} \otimes H_n + B^{(m)}$$

where

$$\{y_0^m...y_T^m\} \overset{DFT}{\Leftrightarrow} Y_m$$

15

$$\{s_{n,0}^m...s_{n,T}^m\} \overset{DFT}{\Leftrightarrow} S_{m,n}$$

$$\{h_{n,0}...h_{n,K-1}\} \overset{DFT}{\Leftrightarrow} H_n$$

Thus, rather than performing a convolution of the channel impulse response h_n with

20   the mth sequence s_{mn} from the nth base transmitter element, the present invention applies a DFT operation to both the received sequence y_n and the various local replicas of the training sequences s_{mn}, and then does a multiplication together of the corresponding frequency bins of the received signal sequence Y_n (being the signal sequence s_n as received through a channel) with the corresponding frequency bins

25   of the replica of the training sequence S_{mn}. Summation (i.e. concatenation) of the individual multiplication products associated with each channel provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel. The process of the present invention generally involves periodic buffering, for example,

31

of frequency bin data and vector results pertaining to channel impulse responses, as will be appreciated by the skilled addressee.

5   In other words, taking an exemplary 128-point Fourier transform of both the received sequence y with the training sequence s yields one hundred and twenty-eight corresponding frequency bins in the frequency domain. Multiplication together of corresponding frequency bins of the received sequence with the same frequency bin of the signal (training) sequence s generates a matrix containing solvable linear equations in H and S. Summation (i.e. concatenation) of the individual multiplication
10   products associated with each channel then provides a composite frequency response for that channel, which composite frequency response can be used in the time domain (through an inverse FFT) to provide a channel impulse response.

It will be appreciated that the channel is assumed constant for the duration of the
15   burst events with respect to slow moving subscriber units, e.g. mobile units having a velocity of less than about ten kilometres per hour (10kph).

In the DFT domain, the kth frequency bin received for the first burst can be expressed as the sum of components from the base stations and their channels:

20
$$Y_1^{(k)} = S_{11}^{(k)} H_1^{(k)} + S_{12}^{(k)} H_2^{(k)} + S_{13}^{(k)} H_3^{(k)} + B_1^{(k)}$$

Similarly, the kth frequency bin for the second burst can be expressed as:

$$Y_2^{(k)} = S_{21}^{(k)} H_1^{(k)} + S_{22}^{(k)} H_2^{(k)} + S_{23}^{(k)} H_3^{(k)} + B_2^{(k)}$$

And, finally, the kth frequency bin for the third burst is:

$$Y_3^{(k)} = S_{31}^{(k)} H_1^{(k)} + S_{32}^{(k)} H_2^{(k)} + S_{33}^{(k)} H_3^{(k)} + B_3^{(k)}$$

25   Rearranging the kth frequency bins provides a set of solvable linear equations in terms of H and S, namely:

$$\begin{pmatrix} Y_1^{(k)} \\ Y_2^{(k)} \\ Y_3^{(k)} \end{pmatrix} = \begin{bmatrix} S_{11}^{(k)} & S_{12}^{(k)} & S_{13}^{(k)} \\ S_{21}^{(k)} & S_{22}^{(k)} & S_{23}^{(k)} \\ S_{31}^{(k)} & S_{32}^{(k)} & S_{33}^{(k)} \end{bmatrix} \begin{pmatrix} H_1^{(k)} \\ H_2^{(k)} \\ H_3^{(k)} \end{pmatrix} + \begin{pmatrix} B_1^{(k)} \\ B_2^{(k)} \\ B_3^{(k)} \end{pmatrix}$$

which can be written in the matrix vector form at each frequency k bin (which for clarity is omitted):

30 $Y=SH+B$

32

It is now possible to define matrices and vectors containing all the transmitter elements and bursts:

$$S = \begin{pmatrix} S_1^{(1)} & S_2^{(1)} & S_3^{(1)} \\ S_1^{(2)} & S_2^{(2)} & S_3^{(2)} \\ S_1^{(3)} & S_2^{(3)} & S_3^{(3)} \end{pmatrix} \qquad H = \begin{pmatrix} H_1 \\ H_2 \\ H_3 \end{pmatrix} \qquad Y = \begin{pmatrix} Y^{(1)} \\ Y^{(2)} \\ Y^{(3)} \end{pmatrix} \qquad B = \begin{pmatrix} B^{(1)} \\ B^{(2)} \\ B^{(3)} \end{pmatrix}$$

5   These equations are solved using a minimum mean square error (MMSE), e.g. Wiener filter, method providing:

$$\hat{H} = \left(S^H S + \mathrm{Cov}(B) \right)^{-1} S^H Y$$

Here $Cov(B)$ is the covariance matrix of the thermal noise, which is often approximated by an identity matrix $\square^2 I$. The equation immediately above can be

10   considered as two entities, namely as a matched filter $S^H Y$ arranged to minimize the signal to noise ratio but not to remove mutual interference between channel estimates, followed by a decoupling matrix $(S^H S + Cov(B))^{-1}$ that removes the coupling between different channels caused by the non-ideal nature of the PN sequences. The inverse DFTs for each radio channel are therefore of the form:

15

$$\begin{pmatrix} H_1^{(0)} \\ H_1^{(1)} \\ H_1^{(2)} \\ \cdot \\ \cdot \\ \cdot \\ H_1^{(z)} \end{pmatrix} \overset{IDFT}{\Rightarrow} \begin{pmatrix} h_1^{(0)} \\ h_1^{(1)} \\ h_1^{(2)} \\ \cdot \\ \cdot \\ \cdot \\ h_1^{(z)} \end{pmatrix}, \quad \begin{pmatrix} H_2^{(0)} \\ H_2^{(1)} \\ H_2^{(2)} \\ \cdot \\ \cdot \\ \cdot \\ H_2^{(z)} \end{pmatrix} \overset{IDFT}{\Rightarrow} \begin{pmatrix} h_2^{(0)} \\ h_2^{(1)} \\ h_2^{(2)} \\ \cdot \\ \cdot \\ \cdot \\ h_2^{(z)} \end{pmatrix}, \ldots$$

where z is the frequency bin (or its time domain equivalent).

The decoupling matrix term $S^H S$ is a matrix product, where S^H is the conjugate transpose of the sequence, i.e. the transposed matrix with all elements complex

20   conjugated. The covariance matrix $<BB^H>$ is the time averaged outer product of the noise vector, or in element-by-element terms:

$$<BB^H> = \begin{pmatrix} <b_1 b_1^*> & <b_1 b_2^*> & <b_1 b_3^*> & \ldots \\ <b_2 b_1^*> & <b_2 b_2^*> & \ldots & \ldots \\ <b_3 b_1^*> & <b_3 b_2^*> & \ldots & \ldots \\ \ldots & \ldots & \ldots & \ldots \end{pmatrix}$$

33

In terms of each channel, concatenation of the numerous frequency bins allows generation of time domain vectors of the channel impulse response for a channel, the time domain vectors being generated from an inverse DFT process within a suitable processing block, such as an ASIC or control processor 150 (of FIG. 1).

It is now possible to extend the basic concept of multiple channel estimation to a situation in which each base station has a plurality of transmit elements; this scenario arises in, for example, systems employing space time coding (STC) or MIMO. It will be appreciated that STC is a collection of techniques for transmitting a number of parallel data streams from multiple transmit antennas with the aim of increasing bit rate for a given total transmitter power. In STC, it will be appreciated that the 2x2 frame structure is expanded to an N-element array, usually having at least four transmit elements. For example, in a four element STC transmitting antenna we might organise the user data in blocks of four symbols and transmit them in different order from the four elements via 16 channels to 4 receiving elements. This is a natural extension of the two element STTD procedure, the main difference from STTD being that multiple receiving elements are essential to get a bandwidth expansion in STC whereas STTD is aimed purely at improving the reliability of the received data and is not concerned with bit rate expansion. Thus STTD will function with a single receive antenna, while STC requires a minimum of two. The principle difference between an STTD arrangement (that is discussed in detail below) and STC is that each component of the block vectors and block matrices (for STTD given below) contains an increasing number of elements (i.e. CIR samples, transmitted data samples and received data samples) consistent with the adopted coding method in each STC derivative.

Referring briefly to FIG. 5, which is generally complementary to FIGs. 1, 3 and 4, there is shown a schematic block diagram of the channel estimation process and architecture 500 of a preferred embodiment of the present invention. The process is based on a two base station (BS1 and BS2) transmission environment having two training bursts. The training sequences $s_x(\square)$ are, in this exemplary case, Gold codes, with individual first 502 and second 504 training sequences transmitted in a first burst 506. The first training sequence 502 is subjected to first channel 508, whereas the second training sequence is subjected to a second channel 510. Both channels are combined at the receiver 512 and are further subject to a first noise component ($N_1(\square)$) 513. In this exemplary case, first and second matched filters 514

34

and 516 provide a time reversed conjugated inversion of the respective first and second training sequences 502 and 504, with outputs from the first and second matched filters 514 and 516 provided to a frequency domain coupler 518 supporting the frequency domain decoupling matrix $(S^H S + Conv(B))^{-1}$. Third 520 and fourth 522

5   training sequences transmitted in a second burst 524 are respectively subjected to first channel 508 and second channel 510. Both channels are combined at the receiver 512 and are further subject to a second noise component 526 (typically different to the first noise component). In this exemplary case, third 528 and fourth 530 matched filters provide a time reversed conjugated inversion of the respective

10   third and fourth training sequences 520 and 522, with outputs from the third and fourth matched filters 528 and 530 provided to the frequency domain coupler 518. The frequency domain coupler then provides channel impulse response vectors $\hat{H}_1(\square)$ and $\hat{H}_1(\square)$ (reference numerals 532 and 534) for the first and second channels, respectively, based on the foregoing point-by-point manipulation of

15   individual (frequency domain) frequency bins.

The preferred embodiment of the present invention therefore shows how accurate channel estimation is possible through the use of Wiener frequency domain MMSE deconvolution combined with frequency domain spatial decoupling matrices, with

20   quasi-orthogonal Gold codes allocated to base stations and their antenna elements.

The principles of the present invention can also be applied to resolve multiple channels by a single training burst; this has particular application in OFDM systems where it may not be possible to transmit Steiner codes because of hardware

25   constraints. The present invention can be modified, in the fashion now shown, to operate with a single arbitrary training sequence yet still resolve multiple channels in a least squares sense. A single training burst situation is algebraically undetermined in the two channel case. If the burst is of length L, there are $2L$ unknowns in the two channels which cannot be determined from L observations at one antenna.

30   However, if there are also $\frac{1}{2}L$ linear constraints associated with each channel impulse response the equations become solvable again. The simplest constraints are that the channel impulse responses $\{h_0 ... h_k\}$ are of finite duration. If $K < \frac{1}{2}L$ then we write the equations as follows. Let S_m be the circulant matrices representing the training bursts and H_m be the IR vectors for channels #1 and #2. However, with a

35   matrix solution which is not circulant, the ability to use FFTs is lost, but resolution of the channels is still possible with use of a matrix of the form:

$$Y = \begin{bmatrix} s_{10} & 0 & s_{20} & 0 \\ s_{11} & s_{10} & s_{21} & s_{20} \\ s_{12} & s_{11} & s_{22} & s_{21} \\ 0 & s_{12} & 0 & s_{22} \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} h_{10} \\ h_{11} \\ h_{20} \\ h_{21} \end{bmatrix}$$

In generality, with respect to two parallel channels using a single training burst, if there are a large number of linear constraints C attached to the channel impulse responses (such as finite durations) then a number of channels can be solved with a

5   single training burst. Algebraically, if the there are N_C constraints on each channel vector of L elements and C channels, the sum of the number of independent degrees of freedom is $C(L - N_C)$. If the length of the training sequence is $N_T \geq C(L - N_C)$, then all of the independent degrees of freedom can be solved; this method is used in Steiner codes to estimate multiple channels using a single training sequence

10

Whilst benefits are attained through use of the channel impulse response estimation scheme described so far (in which the number of processing blocks is equal to the total number of channels to avoid code contention problems), some processing efficiency (i.e. DSP load) associated with Steiner codes is lost since there is a

15   requirement to invert a matrix in each frequency bin whose dimension is equal to the number of channels, i.e. the number of transmitting cells times the number of base antenna elements per cell.

In a further preferred embodiment of the present invention, therefore, the training

20   sequence transmissions of FIG. 4 are implemented as Steiner cyclic training sequence variant; this is shown in FIG. 6 in which successive bursts 602 and 604 from different base stations BS1 and BS2 contain repeated sequence portions of an entire training sequence. This particular embodiment of the present invention offers an optimum combination of frequency domain Wiener filtering and MMSE spatial

25   decoupling for the estimation process. The high accuracy of channel estimation that the combination of techniques offers is a desirable if not essential for coherent null steering and soft hand off in CDMA and OFDM. The sequences $\{s_{11}\}$, $\{s_{21}\}$, $\{s_{12}\}$, $\{s_{22}\}$ are four linearly independent random sequences transmitted in pairs from the two antennas as shown. These sequences transform into the DFT domain as S_{11},

30   S_{21}, S_{12}, S_{11}. Similarly the channel impulse responses are H_1, H_2 in the DFT domain (for the two channels in the exemplary case of a two base station system having

36

single transmit elements). The receiver samples the two received blocks and, via a DFT function, transforms them to the Frequency domain as Y_1 and Y_2. The required guard intervals, which prevent dispersion contamination by unknown data, are already provided in the Steiner sequences and additional, i.e. separate, provisioning

5   of guard periods is not therefore required. The two received blocks are superpositions of convolutions of the transmitted sequences with the channels and are modelled by point by point equations in the DFT domain as follows:

$$\begin{bmatrix} Y_k^1 \\ Y_k^2 \end{bmatrix} = \begin{bmatrix} S_k^{11} & S_k^{12} \\ S_k^{21} & S_k^{22} \end{bmatrix} \begin{bmatrix} H_k^1 \\ H_k^2 \end{bmatrix} + \begin{bmatrix} N_k^1 \\ N_k^1 \end{bmatrix} \quad k = 0...,2^n - 1$$

or Y(k) = S(k).H(k)+(N(k) which has a least squares solution

10    $$\hat{H}(k)_{LS} = \left(S(k)^H S(k) \right)^{-1} S(k)^H Y(k) \qquad k = 0 \cdots 2^n - 1$$

in each frequency bin and can be shown to have a MMSE solution

$$\hat{H}(k)_{MMSE} = \left(S(k)^H R_N^{-1} S(k) + R_H^{-1} \right)^{-1} S(k)^H R_N^{-1} Y(k)$$

Where R_N is the noise covariance and R_H is the channel covariance. In general, for N_{TX} transmitters, there are N_{TX} channels to estimate and N_{TX} blocks of pilot or

15   training sequences are required for a solution.

Since the S matrix in the basic equation above has no particular orthogonality properties, matched filtering is not advisable. Instead, if the following MMSE matrix operation is performed at each frequency; the result combines spatial decoupling

20   with a Wiener frequency weighting function and overall gives an MMSE solution. It requires a 2×2 matrix inversion in each frequency bin for two cell operation (with each base station having a single transmit element):

$$\begin{bmatrix} \hat{H}_k^1 \\ \hat{H}_k^2 \end{bmatrix} = \left(\begin{bmatrix} S_k^{11} & S_k^{12} \\ S_k^{21} & S_k^{22} \end{bmatrix} + \sigma^2 \mathbf{I} \right)^{-1} \begin{bmatrix} Y_k^1 \\ Y_k^2 \end{bmatrix} \quad k = 0...,2^n - 1$$

By combining Steiner cyclic pilot rotation for transmissions from the transmit

25   elements of each base station according to the present invention, interference-free mobiles near a centre of the cell are allowed to resolve their channels in the usual (conventional) way. At the cell periphery where interference becomes significant, the training sequences need to be expanded to provide enough degrees of freedom such that the simultaneous equations for *all* of the channels become algebraically

30   determinate; a repeating pattern of the Steiner PN sequences allows this to occur.

Using the MMSE formulas described in the Klein paper identified above, we get the following alternative forms of the above MMSE equation

$$\hat{H}(k)_{MMSE} = R_H S(k)^H \left(S(k) R_H S(k)^H + R_N \right)^{-1} Y(k)$$
$$= \left[I + (R_H S(k)^H R_N^{-1} S(k))^{-1} \right]^{-1} \hat{H}(k)_{LS}$$

5   The last equation shows that the LS estimate above can be calculated first and then transformed into an MMSE one. If we set $R_N = \sigma_N^2 I$ and $R_H = \sigma_H^2 I$ in MMSE equation above, then

$$\hat{H}(k)_{MMSE} = \left(S(k)^H S(k) + \frac{\sigma_N^2}{\sigma_H^2} I \right)^{-1} S(k)^H Y(k)$$

However R_N is modified by the dispersive channel. It is possible to estimate R_N
10   on-line since the auto- and cross-covariance statistics of the received pilots and data are effectively identical. The covariance of the noise vector is block diagonal since the statistics of consecutive blocks are independent and for example in an $N_{TX} = 3$ case, R_N has the form

$$R_N(k) = \begin{bmatrix} R_{YY}(k) & 0 & 0 \\ 0 & R_{YY}(k) & 0 \\ 0 & 0 & R_{YY}(k) \end{bmatrix}$$

15   where $R_{YY}(k)$ is the estimated signal covariance measured across the receiver elements at frequency k. Note that R_{YY} contains a component due to the pilot signal itself (at −10 dB) as well as the target interfering data signals which can be cancelled as necessary.

Both the matrix solutions of the least squares estimate and the MMSE estimate
20   compensate for channel estimation errors caused by a pilot signal which has a non-uniform power spectrum and a non-zero autocorrelation. When these equations are solved instead of a simple correlation of the form:-

38

$$\hat{H}_{i1}(k) = F_1(k)^*\big(H_{i1}(k)F_1(k) + H_{i2}(k)F_2(k)\big)$$

$$= \big|F_1(k)\big|^{2^*} H_{i1}(k) + F_1(k)F_2(k)H_{i2}(k)$$

$$\hat{H}_{i2}(k) = F_2(k)^*\big(H_{i1}(k)F_1(k) + H_{i2}(k)F_2(k)\big)$$

$$= F_2(k)^* F_1(k)H_{i2}(k) + \big|F_2(k)\big|^2 H_{i2}(k)$$

where $F(k)$ is the pilot or training sequence (equivalent to $S(k)$ above), the solutions are consistent in the sense that as the SNR or the pilot sequence duration increases to a large value, the estimated channel taps converge asymptotically on their true values. This is not true for the simple correlation.

While no MIMO radio standards presently exist which allow for such multiple pilot sequences, existing pilot blocks can be subdivided into smaller sections at the terminal to generate the necessary equations so the dual pilot block method is compatible with Release 5 pilot arrangements (3G Specification TSG TS25.211: "Physical channels and mapping of transport channel on physical channels (FDD)", Release '99, Oct 1999, ftp://ftp.3gpp.org/Specs/2000-06/R1999/25_series/25211-330.zip). Either the estimator can use two consecutive slots in the frame to estimate the channel, getting uncorrelated scrambling codes, or a single slot can be divided into two parts. These choices involve trades-offs between Doppler and noise sensitivity. If there are $M>2$ TX antennas, the number of sequences required to algebraically determine the channels increases to M per antenna, a total of M^2 altogether, transmitted in M blocks.

In the exemplary case where a receiver observes six sequences, as shown in FIG. 7, the result in the Fourier domain is that there are sufficient degrees of freedom to solve for six channels $\{h_1\}$ to $\{h_6\}$. It should be noted that in FIG. 7, the introduction of sign reversals allows for a distinction to be drawn between codes in successive bursts, with the mechanism of sign reversal therefore reducing the overall number of Steiner codes required to implement a system according to the present invention. The equations relating the received signal Y to the sequences U, V, W and the channels H in the frequency domain are:

$$
\begin{bmatrix} Y_1^{(k)} \\ Y_2^{(k)} \\ Y_3^{(k)} \\ Y_4^{(k)} \\ Y_5^{(k)} \\ Y_6^{(k)} \end{bmatrix} = \begin{bmatrix} U_1^{(k)} & U_2^{(k)} & V_1^{(k)} & V_2^{(k)} & W_1^{(k)} & W_2^{(k)} \\ U_2^{(k)} & U_1^{(k)} & V_2^{(k)} & V_1^{(k)} & W_2^{(k)} & W_1^{(k)} \\ U_1^{(k)} & U_2^{(k)} & -V_1^{(k)} & -V_2^{(k)} & W_1^{(k)} & W_2^{(k)} \\ U_1^{(k)} & U_2^{(k)} & -V_2^{(k)} & -V_1^{(k)} & W_2^{(k)} & W_1^{(k)} \\ U_1^{(k)} & U_2^{(k)} & V_1^{(k)} & V_2^{(k)} & -W_1^{(k)} & -W_2^{(k)} \\ U_1^{(k)} & U_2^{(k)} & V_2^{(k)} & V_1^{(k)} & -W_2^{(k)} & -W_1^{(k)} \end{bmatrix} \begin{bmatrix} H_1^{(k)} \\ H_2^{(k)} \\ H_3^{(k)} \\ H_4^{(k)} \\ H_5^{(k)} \\ H_6^{(k)} \end{bmatrix} + \begin{bmatrix} N_1^{(k)} \\ N_2^{(k)} \\ N_3^{(k)} \\ N_4^{(k)} \\ N_5^{(k)} \\ N_6^{(k)} \end{bmatrix} \quad k = 0...2^n - 1
$$

However this is an inefficient algebraically redundant representation that can be simplified as follows. At the mobile terminal, for example, the signal from a given antenna at a base station is the circular convolution of the PN sequence {s} with the channel impulse response {h}:

$$
y_k = \sum_{j=0}^{N-1} s_{k-j\,\mathrm{MOD}\,N} h_j \quad k = 0...2^n - 1
$$

The cyclic shift of the PN sequence for second base element in the Steiner method can be replaced by a reverse shift of the channel impulse response to give exactly the same signal at the subscriber unit:

$$
y_k = \sum_{j=0}^{N-1} s_{k-j-N/2\,\mathrm{MOD}\,N} h_j
$$
$$
= \sum_{j=0}^{N-1} s_{k-j} h_{j+N/2\,\mathrm{MOD}\,N}
$$

Cyclic shifting of channel impulse responses h_2, h_4 and h_6 (of FIG. 7) takes into account the reverse shift as detailed immediately above with the effect that the six sequences received at the antenna of the subscriber (or other addressed unit) are $\underline{Y}_1$, $\underline{Y}_2$, $\underline{Y}_3$, $\underline{Y}_4$, $-\underline{Y}_5$ and $-\underline{Y}_6$. Moreover, the six sequences are obtained from a repetition of only three Steiner codes that are individually and uniquely assigned to each base station, i.e. both transmit elements send identical sequences in the same burst, with the subscriber unit generating a cyclically shifted channel impulse response $\underline{h}'_x$ for one of the two radiating elements of the base station's transmit array. This reverse shift manipulation results in only three different PN sequences being utilised in this particular embodiment, with the channel impulse responses of FIG. 7 taking the form ($\underline{h}_1 + \underline{h}'_2$), ($\underline{h}_3 + \underline{h}'_4$) and ($\underline{h}_5 + \underline{h}'_6$) to provide the received sequence $\underline{Y}_1$, $\underline{Y}_2$, $\underline{Y}_3$, $\underline{Y}_4$, $-\underline{Y}_5$ and $-\underline{Y}_6$. The determinant of the sign change matrix of this solution is 4, as shown below:

40

$$\begin{vmatrix} 1 & 1 & 1 \\ 1 & -1 & 1 \\ 1 & 1 & -1 \end{vmatrix} = 4$$

Writing

$$\mathbf{y}_1 = \begin{bmatrix} y_1 \\ y_2 \end{bmatrix} \quad \mathbf{y}_2 = \begin{bmatrix} y_3 \\ y_4 \end{bmatrix} \quad \mathbf{y}_3 = \begin{bmatrix} y_5 \\ y_6 \end{bmatrix}$$

$$\mathbf{h}_1 = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} \quad \mathbf{h}_2 = \begin{bmatrix} h_3 \\ h_4 \end{bmatrix} \quad \mathbf{h}_3 = \begin{bmatrix} h_5 \\ h_6 \end{bmatrix}$$

$$\mathbf{u}_1 = \begin{bmatrix} u_1 \\ u_2 \end{bmatrix} \quad \mathbf{v}_1 = \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \quad \mathbf{w}_1 = \begin{bmatrix} w_1 \\ w_2 \end{bmatrix}$$

with Fourier transforms:

$$\mathbf{y}_1 \Rightarrow \mathbf{Y}_1 \quad \mathbf{y}_2 \Rightarrow \mathbf{Y}_2 \quad \mathbf{y}_3 \Rightarrow \mathbf{Y}_3$$
$$\mathbf{h}_1 \Rightarrow \mathbf{H}_1 \quad \mathbf{h}_2 \Rightarrow \mathbf{H}_2 \quad \mathbf{h}_3 \Rightarrow \mathbf{H}_3$$
$$\mathbf{u} \Rightarrow \mathbf{U} \quad \mathbf{v} \Rightarrow \mathbf{V} \quad \mathbf{w} \Rightarrow \mathbf{W}$$

5

then a reduced set of more easily solvable equations in the DFT domain is obtained, namely:

$$\begin{bmatrix} \mathbf{Y}_1^{(k)} \\ \mathbf{Y}_2^{(k)} \\ \mathbf{Y}_3^{(k)} \end{bmatrix} = \begin{bmatrix} \mathbf{U}^{(k)} & \mathbf{V}^{(k)} & \mathbf{W}^{(k)} \\ \mathbf{U}^{(k)} & -\mathbf{V}^{(k)} & \mathbf{W}^{(k)} \\ \mathbf{U}^{(k)} & \mathbf{V}^{(k)} & -\mathbf{W}^{(k)} \end{bmatrix} \begin{bmatrix} \mathbf{H}_1^{(k)} \\ \mathbf{H}_2^{(k)} \\ \mathbf{H}_3^{(k)} \end{bmatrix} \quad k = 0...2^n - 1$$

Provided a resultant matrix in this set of equations is well conditioned and invertible,
10   then the set of channel impulse responses can be solved. A sampling design for the repetitions that offers near-orthogonality between repeated sequences is, however, required, such that different bases can be resolved without ill-conditioning of the equations. Such designs, as used for space-time codes, are readily available in statistics literature.

15

By employing the preferred embodiment of the present invention where Steiner codes supplement to point-by-point DFT multiplication mechanism, accuracy in estimated channel impulse responses is limited by channel noise rather than imperfection in the codes.

20

The preferred embodiment that uses Steiner codes for m base antenna elements can be used in multiple cells and MMSE channel estimation is obtainable for a

number n of bases giving a total of nm downlink channels. Different cells will use a different basic PN code, whereas the elements within a single cell will use the Steiner cyclic offsets of a single PN code. The extra workload (processing overhead) involved is the inversion of an $n \times n$ matrix in each frequency bin. The inclusion of the

5 use of Steiner codes to supplement the basic inventive concept of Wiener frequency domain MMSE deconvolution combined with frequency domain spatial decoupling matrices results in the possibility of allocating only a single training sequence to each base station provided that the training sequence is of sufficient length to encompass all the multiple time-translated channel impulse responses.

10

Embodiments of the invention may be implemented as computer program code encoded on a computer program product for use with a computer system. It is expected that such a computer program product may be distributed as a removable medium with accompanying printed or electronic documentation (e.g. shrink-

15 wrapped software), preloaded with a computer system or distributed from a server or electronic bulletin board over a network (e.g. the Internet or World Wide Web). A series of computer instructions can therefore either be fixed on a tangible medium or fixed in a computer data signal embodied in a carrier wave that is transmittable to a computer system using wireline or wireless transmission techniques. For example,

20 the present invention can find application as downloadable code in a software radio. The removable (i.e. tangible) medium may be a computer readable media, such as a diskette, CD-ROM, DVD-ROM or RAM, fixed disk, magneto-optical disks, ROMs, flash memory or magnetic or optical cards. The series of computer instructions embodies all or part of the functionality previously described herein with respect to

25 the system.

Software embodiments of the invention may be implemented in any conventional computer programming language. For example, preferred embodiments may be implemented in a procedural programming language (e.g. "C") or an object oriented

30 programming language (e.g. "C++").

Although the preferred operating method is realised by general or specific-purpose processor or logic circuits programmed with suitable machine-executable instructions, hardware components may possibly be used to implement certain

35 features of the present invention. Of course, the present invention may be performed by a combination of hardware and software.

It will, of course, be appreciated that the above description has been given by way of example only and that modifications in detail may be made within the scope of the present invention. For example, while the preferred embodiment is has been

5   described in relation to a multi-cast system in which each base station has a single transmit element, the present invention is clearly applicable to multi-element transmitters and to single isolated cells supporting transmissions from a single base station having a transmit array. Furthermore, while the preferred embodiment has been described in the context of a CDMA system, the present invention is generally

10   applicable to any dispersive or imperfect channel that acts to corrupt symbols, including systems employing TDD and FDM techniques and other types of links, such as digital subscriber lines (DSL) and all digital transmission systems.

Indeed, although the technique of the preferred embodiment of the present invention

15   has been described for a CDMA application where the number of chips in the training burst is easily identified, the principles of the present invention apply to OFDM, wavelets (e.g. Daubechies's wavelets) and any similar orthogonal or near-orthogonal signaling waveforms. With specific regard to an OFDM signal, such an OFDM signal is treated as a single wide-band modulated carrier in the time domain

20   and an estimate of the time domain channel impulse response is made in an identical way to that described in the preferred embodiment. If a constraint is made that the duration of the impulse responses is small, the equations become algebraically soluble (or *observable* in the context of control theory jargon) and in this way a large number of parallel channels can be estimated using a single training

25   burst.

If the duration of the OFDM training burst is N_S sub-carrier symbols and there are N_F sub-carriers then the burst effectively has N_S*N_F samples or chips which is quite a substantial number. In practice, this means that multiple channel estimation in

30   OFDM might not be too difficult. The lower limit for the duration of the training sequences would be that it must be capable of supporting an orthogonal, or at least a linearly independent, set of codes for say m base station antennas seen from the terminal (in one or more cells). Superficially, this would require m full sub-carrier slots and for HSD with 150 µs slots we need a duration of $150m$ µs. However the

35   wideband methods described here only require that the duration of the training sequence is $\geq m$ times the duration of the channel impulse response so, for a

43

channel with τ μs dispersion and signal bandwidth B MHz, the duration of the training sequence need only be τBm samples. For example, if the signal bandwidth is 6 MHz, the channel dispersion is 5μs and there are seven base station antennas with channels to be determined, the minimum duration of the training sequence is

5   210 samples or $6 \times 5 \times 7/6 = 35\mu$s, i.e. only 1/5 of a nominal sub-carrier symbol. In practice, however, this duration would need to be increased significantly (and preferably about ten-fold or so) to improve estimation accuracy.

In transmission bursts from base stations having single transmit elements, cyclic

10   prefixes could be replaced by blank (zero) carriers to provide isolation between burst sequences.

A technique for improving channel estimations (obtained using the above MMSE multivariate technique or in another way) using a long-term power-delay profile

15   (PDP), is now described.

The ITU has defined channel impulse responses for 3G mobiles which are truncated at a few microseconds:

Pedestrian and Outdoor to Indoor; A and B: 0.41 and 3.7 μs respectively

20   (1.6 and 14.2 chips)

Vehicular; A and B: 2.51 and 20 μs respectively (9.64 and 76.8 chips)

and the relative multipath powers are quite small at these extremes; typically around −20 dB.   In any type of estimation it is advisable to estimate as few parameters as possible since introducing unnecessary degrees of freedom is not

25   necessarily a neutral procedure.   The extra parameters may be given phantom values created from the noise and interference which degrade performance.   Also, an equalizer, given enough degrees of freedom, can condition the random noise in a finite window to become highly correlated with the pilot signal and reduce the output MS pilot error but this corrupts the desired signal even more and increases

30   BER.

A suitable procedure using the matrix-vector algebra presented above, can be derived. It can be shown that:

44

$$\hat{H}_{MMSE} = \left(P^H R_{NN}^{-1} P + R_{HH}^{-1}\right)^{-1} P^H \left(R_{NN}^{-1} X + R_{HH}^{-1} H_0\right)$$

Where $\hat{H}$ is the channel matrix, H_0 is a prior mean value estimate of an element H, R_{HH} is the channel covariance and R_{NN} is the noise covariance.

5

Both the H covariance R_{HH} and H_0 can be used as a vehicle to inject prior knowledge into the estimation. Typically we have no prior knowledge of the complex values of the channel taps in the impulse response but know their likely variance, either from a general statistical knowledge of the channel and its likely dispersion or from long term observations *in situ*. If R_{HH} is diagonal, i.e. the channel taps are uncorrelated, its diagonal values represent the expected power-delay profile of the channel. Fig. 8 shows the power-delay profile of the ITU vehicular-B channel. Such a profile could be estimated by the UE over a long term and used as diagonal entries in R_{HH}.

15

As a simpler approximation, working with a DFT block size of say 4096 and the Ped-A model of duration 4μs, we know that channel taps 0 to 15 are likely to have $|h_k|$ values up to unity while taps 16 to 4095 are zero. Thus we can partition the R_{HH} matrix into four sections as shown to reflect these expected values

20
$$R_{HH} = \begin{bmatrix} I_{16\times16} & 0 \\ 0 & 10^{-6} I_{4080\times4080} \end{bmatrix}$$

Ideally the lower 4080×4080 submatrix should be set to exactly zero but then the overall problem becomes singular as it stands and is difficult to solve. Inversion of the lower submatrix of R_{NN}^{-1} gives large (10^6) diagonal values in the solution to te MMSE equation whose effect is to swamp values in rows and columns 16 to 4095 in the matrix $\left(P^H R_{NN}^{-1} P + R_{HH}^{-1}\right)$ and ultimately to set them to zero in the inverse$\left(P^H R_{NN}^{-1} P + R_{HH}^{-1}\right)^{-1}$. Effectively the solution is projected onto a subspace orthogonal to columns 16 to 4095 of the matrix $P^H R_{NN}^{-1} P$.

Examining what is left:

$$\begin{bmatrix} \hat{H}_{MMSE}^{TRUNC} \\ 0 \\ \vdots \end{bmatrix} = \begin{bmatrix} \left(P^H R_{NN}^{-1} P\right)_{16 \times 16}^{-1} & 0 & \cdots \\ 0 & 0 & \cdots \\ \vdots & \vdots & \ddots \end{bmatrix} \begin{bmatrix} \left(P^H X\right)_{16 \times 1} \\ 0 \\ \vdots \end{bmatrix}$$

we have a truncated matrix $P^H R_{NN}^{-1} P$ which is Toeplitz rather than circulant and so is incompatible with FFT's. In order to further reduce the complexity of the problem, an approximate solution can be obtained in Fourier space by extrapolation of the

5   Toeplitz matrix to a circulant matrix. Writing $T = P^H R_{NN}^{-1} P$, the diagonals of T are extrapolated right down the leading diagonals of a large (up to 4096×4096) empty matrix to produce a circulant (Figure 9)

In the FFT method the equivalent sequence of operations can be found by

10   introducing an DFT matrix F, where the product FF^H is an identity matrix. The MMSE equation then becomes

$$F\hat{H}_{MMSE} = F\left(F^H P^H FF^H R_{NN}^{-1} FF^H PF + F^H R_{HH}^{-1} F\right)^{-1} F^H FP^H FF^H R_{NN}^{-1} FF^H X$$

For circulant P and diagonal R_{NN} we have $F^H P^H F = \Lambda_P$ $F^H R_{NN}^{-1} F = \sigma^2 I$ and the last equation becomes

15    $$F\hat{H}_{MMSE} = F\left(\Lambda_P^H \sigma^2 I \Lambda_P + F^H R_{HH}^{-1} F\right)^{-1} F^H \Lambda_P^H \sigma^2 I \Lambda_P F^H X$$

Everything in this equation can be identified as a simple FFT method except for the matrix R_{HH}. This matrix

is to approximate the equivalent of $P^H R_{NN}^{-1} P$,

$$P^H R_{NN}^{-1} P \Rightarrow A(\omega_k) = \frac{|P(\omega_k)|^2}{|R(\omega_k)|}$$

20    $A(\omega_k)$ is then IFFT transformed to the time domain autocorrelation function, $a(\tau)$, this is truncated at $\pm T_{max}$, the maximum duration of the channel impulse response, then FFT transformed back to the frequency domain. This spectrum is then equivalent to the time-truncated circulant matrix above. In the case of a MIMO situation the matrix $P^H R_{NN}^{-1} P$ is block circulant rather than a simple circulant but this

25   can easily be accommodated within the same general solution.

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A second estimation method uses a non-informative prior, a LS estimation procedure, and inserts known channel tap variances after the H vector has been obtained. The reasoning behind this method is that if the covariance of the

5   estimated taps in a derivation of the MMSE equation:

$$\hat{H}_{MMSE} = \left(P^H P + \frac{\sigma_N^2}{\sigma_H^2} I \right)^{-1} P^H X \quad \text{(assuming } R_{HH} \text{ and } R_{NN} \text{ are both identity matrices)}$$

is almost diagonal then it should not matter whether the H covariance is used within or after the equation. For example, setting $\sigma_H{}^2 = 1$ for all taps initially (a reasonable guess for channel taps known to be less than unity in value) then we get a first

10   estimate H_1:

$$\hat{H}_1 = \left(P^H R_{NN}^{-1} P + R_{HH}^{-1} \right)^{-1} P^H R_{NN}^{-1} X \approx \left(\Lambda^{-1} + \sigma_H^{-1} 1 \right)^{-1} P^H R_{NN}^{-1} X$$

then we can approximate the full MMSE procedure with a known R_{HH} by

$$\hat{H}_{MMSE} = \left(\Lambda^{-1} + R_{HH}^{-1} \right)^{-1} \left(\Lambda^{-1} + \sigma_H^{-2} I \right) \hat{H}_1$$

and letting $\Lambda = \lambda I$ corresponding to an uncorrelated thermal noise with

15   near-orthogonal pilots and $R_{HH} = \text{Diag}(\rho_1, \rho_2, \ldots)$ then

$$\hat{H}_{MMSE} = \text{Diag}\left(\frac{\lambda + \rho_k}{\lambda \rho_k} \right)^{-1} \text{Diag}\left(\frac{\lambda + \sigma^2}{\lambda \sigma^2} \right) \hat{H}_1$$

$$= \text{Diag}\left(\frac{\rho_k}{\sigma^2} \cdot \frac{\lambda + \sigma^2}{\lambda + \rho_k} \right) \hat{H}_1$$

The action of this equation is to suppress tap estimates in the vector H_1 whose a--priori standard deviation values ρ_k lie below the estimation standard deviation λ of the LS estimator (Figure 10), independently of the actual observed tap values. The

20   ρ_k values are determined by long term observations of channel tap estimates on a lengthy sequence of slots and are parameter deviations rather than estimation accuracies while the λ values derive from the pilot sequences and knowledge of the interference source (thermal noise + superimposed data).

25   Thus this Bayesian technique is not a linear channel-tap predictor in the "long range prediction" sense suggested, for example in, R1-01-0025: "On the need of Long-

47

Range Prediction (LRP) of channel estimation in HSDPA and text proposal",
Wiscom, 15-19 Jan. 2001, rather it is a method for eliminating insignificant channel
taps which are historically of small amplitude in relation to accuracy. Faced by an
unknown channel which could be the Vehicular-B IU channel of up to 85 taps in
5   duration with six active reflections in unknown locations or the Ped-B ITU channel of
only 1.6 chips duration, the long term variance estimation can reduce the channel
model to a small number of taps in either case and reduce the aggregate channel
estimate error variance several-fold relative to a model with a large number of taps.

10   With reference to Figure 11, a further improvement (on an estimation produced
using one of or a combination of the techniques described above, or of another
channel estimation technique) is now described.

As explained above, the use of orthogonal pilot codes in m transmit antennas is
15   insufficient to completely differentiate their transmissions at the receiver in a
dispersive channel but the problem is substantially solved by the use of m-fold
multiple versions of the training sequences giving a determinate algebraic solution
for the m channel impulse responses to each receiver element. However, in the
presence of user data on the transmissions further improvements on that technique
20   can be achieved since the data superimposed on the pilot signals is also
deorthogonalized in a dispersive channel and the data noise makes the channel
estimates noisy. The data is unknown at the receiver so it is not possible to write
out equations to remove was done for the pilots/training sequences in the technique
above.

25

For the 3GPP FDD standard, the CPICH codes are modulated repetitions of the
256-chip Walsh code #0, which is all ones, while the data modulates orthogonal
16-chip Walsh codes drawn from the orthogonal variable spreading factor (OVSF)
tree. Orthogonality requires that the user Walsh codes are codes other than 16-chip
30   code #0 which is 16 ones, so there are only 15 codes available for user data.

48

Given this problem, iterative improvement of the channel coefficients is proposed as an effective method for improving the receiver performance to a condition very close to optimum.

5   The goal for optimising the channel coefficients in a dispersive channel is to achieve zero cross coupling of the demodulated user-data and pilot symbols and a good equivalent metric is the mean square error of the demodulated pilot symbols. The basic system is shown in Figure 11 which shows a system for iteration of a Fourier domain solution by minimising mean square error (MSE) of the received pilots
10   against a local 'ideal' copy as explained below.

The instinctive implementation demands a large DSP load since the Fourier transforms have to be redone at every iteration. High DSP load is undesirable. To optimise the receiver it is has been found to be sufficient to optimise reception of the pilots by demodulating them and minimising the mean square error of their
15   associated "symbols". The "pilot error" signal could be at the chip level or after despreading by a factor of 16 to match the data.

Let the summed mean square (MS) error θ of the two pilots symbol sequences at the receiver be functionally dependent on the estimated channel impulse responses by a function of the form θ ($\underline{h}$) where $\underline{h}$ is the vector of Channel Impulse Responses
20   (ChIR) for MN channels (M transmit antenna elements and N receive antenna elements). The function θ ($\underline{h}$) is not known at the outset but by using the estimated channels in an earlier process such as one described above, we start out fairly close to a minimum of θ. If the equalizer were a linear function of the channel coefficients then the MS error at the output would be a quadratic function of h having the general
25   form

$$\theta(h) = (h - h_{OPT})^H M (h - h_{OPT}) + c$$

where c is the minimum error, M is a matrix representing cross coupling between the channel components, h contains all channel impulse responses stacked vertically into a single vector and h_{OPT} is the optimum channel estimate.

30   The equaliser derived from h is actually close to a channel whitener so its frequency spectrum approximates a reciprocal of the spectrum of h. Therefore θ is not

actually quadratic function of h as suggested above. An example for a six-tap channel is shown in Figure 12

General methods for exploring non-linear responses are known, however, for small perturbations about the optimum h_{OPT} it appears that the quadratic equation above is
5   a good approximation to $\theta\,(\underline{h})$. The simulations have minimised $\theta\,(\underline{h})$ by symmetric perturbations of the vector h. It has been found to be sufficient to do this element by element as there is little coupling between taps. For MN physical channels each with a K-tap impulse response, the number of variables is $2KMN$, treating the in-phase and quadrature elements of h as independent variables, thus we have this
10   many variables to optimise. The minimum for the nth element h_n is found by perturbing of h_n by $\pm\delta$ and fitting a parabolic function to the three values of θ as shown in Figure 13.

$$h + \delta_n = \begin{bmatrix} h_0 \\ h_1 \\ h_2 + \delta_2 \\ \\ h_{N-1} \end{bmatrix}$$

15   We can write $\theta = ax^2 + bx + c$ on this line with $\delta = 1$ and we have

$$d_2 = \theta(x_n + \delta_n) - 2\theta(x_n) + \theta(x_n - \delta_n) = 2a$$
$$d_1 = \theta(x_n + \delta_n) - \theta(x_n - \delta_n) = 2b$$
$$c = \theta(x_n)$$

and the minimum of the parabola lies at $x'_n = x_n - \delta\dfrac{d_1}{2d_2}$. Thus after measuring the

error at the three points the associated channel impulse response element is reset to x'_n before moving to the next element.

20

Figure 14 shows the iterative improvement of the equalizer for an exemplary 2×2 MIMO channel. The four physical channels each have two tap channel impulse responses with coefficients drawn independently from a complex Gaussian

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distribution. The data uses BPSK modulation in a block of 1024 samples and the system is noise-free. There are 16 real-valued channel coefficients (= 8 complex numbers) which describe the channel and these have been optimised in turn in 4 cycles.

5   The iterative channel adjustment of the last section is formalized in Figure 15 which shows the long-term or a-priori channel power-delay profile (PDP) 1500, a sparse complex channel impulse response estimate $\hat{g}_k$ 1510, generation of a frequency domain Wiener equaliser from $\hat{g}_k$ 1520, equalization of the channel in the frequency domain 1530 and measurement of the mean square difference θ 1540

10   between the equalized pilot signal from the channel 1550 and the local reference 1560. A feedback loop 1570 from the latter is used to optimise $\hat{g}_k$ by minimizing θ.

Note that as the equalizer is improved in this iterative method, so the orthogonality between the data and pilot signals is progressively restored and the "user data noise" superimposed on the pilots at a relative level of +10 dB, which dominates the

15   channel estimation noise in a dispersive channel, starts to decrease and finally becomes insignificant. The general method easily extends to finding MIMO channel FFT equalizers and this was done for the 2:2 channel results of Figure 14 where there was a >1000-fold improvement in MS accuracy during adaptation from an initial least-squares channel estimate.

20   The long term power-time profile (PDP) of the channel can use information from a number of sources. Firstly there is knowledge of the maximum time delay likely in a given channel, secondly there are ChIR estimates done with a simplified estimator possible direct correlation, thirdly old ChIR estimates from previous slots and fourthly the ACF of the composite signal, which includes the data part, can be

25   analysed to get a ChIR. The ACF is an added result from all the TX antennas but it is expected that in a typical MIMO situation the locations of multipaths will be very similar in all the channels and the composite TX multipaths will give good prior information.

30   It should be understood that reference to a base station refers to serving infrastructure irrespective of the form of air-interface adopted by the exemplary CDMA–based architecture of the preferred embodiment. In other words, the term base station should be construed broadly and, where necessary, to include

references to communication relay devices or even mobile stations. While the main applications are in channel estimation on a downlink for CDMA and OFDM in order to apply coherent soft hand-off and interference cancellation in locations where two or more base stations are at similar ranges and the weaker base stations, if ignored,

5   cause mutual interference, the preferred embodiments are not intended to be limiting to these specific arrangements.

In summary, multiple Steiner codes are transmitted as bursts (s_{11}, s_{12},... s_{33}, 560, 524) from multiple base stations (182, 184, 186) having one or more transmit

10   elements (174, 176, 178, 180), with successive bursts providing an extended training sequence for use in channel estimation at an addressed unit (172), such as a mobile handset. Accurate channel estimation is possible through the use of Wiener frequency domain MMSE deconvolution (518) combined with frequency domain spatial decoupling matrices, with quasi-orthogonal pseudo-noise sequences

15   (502, 504, 520, 522) allocated to base stations and their antenna elements. The use of Steiner codes to supplement Wiener frequency domain MMSE deconvolution and frequency domain spatial decoupling results in the possibility of allocating only a single training sequence to each base station provided that the training sequence is of sufficient length to encompass all multiple time-translated channel impulse

20   responses (H). Estimates may be refined iteratively by minimising the MS error of demodulated pilot symbols. Estimates may also be refined by removing taps from the impulse response which are insignificant based on a relatively long-term power-delay profile for the channel.

We claim:

1.   A method of determining channel impulse responses of a plurality of
5   channels to a communication device, the method comprising:

        performing transform operations on both a replica of a signal sequence s_n
and a received training sequence y_n received by the communication device in at
least one burst, the received training sequence y_n being the signal sequence as
received through a channel, the transform operations arranged to generate a
10   multiplicity of signal sequence frequency bins and a multiplicity of training sequence
frequency bins;

        performing point-by-point operations between corresponding signal
sequence frequency bins and training sequence frequency bins; and

        concatenating the point-by-point operations associated with the channel to
15   provide a composite frequency response for the channel, the composite frequency
response allowing, in the time domain, generation of the channel impulse response
for the channel.

2.   The method according to claim 1, further comprising using a set of matrix
20   operations in the frequency domain to resolve channels to the communication
device from multiple transmitting stations, the matrix operations providing solvable
linear equations for the channel impulse response H and the signal sequence S
expressible in a matrix-vector form at each frequency bin in terms of the received
training sequence Y.
25

3.   The method according to claim 2, further comprising solving the linear
equations using a minimum mean square error (MMSE) estimation technique.

4.   A method of determining channel impulse responses of channels incident to
30   a communication device, the method comprising: and

        transmitting multiple quasi-orthogonal pseudo-noise sequences as bursts
from multiple base stations each having at least one transmit element, successive
bursts providing an extended training sequence for use in channel estimation at the
communication device;

53

applying a Wiener frequency domain MMSE deconvolution with frequency domain spatial decoupling matrices to generate channel impulse response estimates for the channels.

5   5.   A communication device having a receiver coupled, in use, to receive a plurality of channels supporting a signal sequence y_n and training sequence bursts, the communication device having:

a signal processing platform to perform transform operations on both a replica of a signal sequence s_n and a received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins;

the signal processing platform arranged to perform point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency bins; and

the signal processing platform further arranged to concatenate the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel.

6.   The communication device according to claim 5, wherein the signal processing platform operates to establish a set of matrix operations in the frequency domain to resolve channels to the communication device from multiple transmitting stations, the matrix operations providing solvable linear equations for the channel impulse response H and the training sequence S expressible in a matrix-vector form at each frequency bin.

7.   The communication device according to claim 6, wherein the signal processing platform solves the linear equations using a minimum mean square error (MMSE) estimation technique.

8.   A communication receiver comprising:

means for receiving, in use, multiple quasi-orthogonal pseudo-noise sequences as bursts from multiple base stations each having at least one transmit

element, successive bursts providing an extended training sequence for use in channel estimation at the communication receiver; and

means for applying a Wiener frequency domain MMSE deconvolution with frequency domain spatial decoupling matrices to generate channel impulse
5   response estimates for the channels.

9.   A base station of a communication system, the base station comprising:

a transmitter chain arranged to transmit multiple quasi-orthogonal pseudo-noise sequences as training bursts s_n from at least one transmit element and further
10 arranged to transmit a signal sequence y_n, successive training bursts providing an extended training sequence for use in channel estimation at a communication device of the communication system, the transmitter chain substantially simultaneously transmitting, in use, different training bursts from each of the at least one transmit element, each training burst having a length at least as long as a
15 maximum channel duration in the communication system multiplied by a number corresponding to a plurality of channels to the communication device, the extended training sequence and the signal sequence y_n providing a resolution mechanism to the communication device allowing the communication device to resolving the plurality of channels to recover associated channel impulse responses H for each
20 channel by solving an algebraic matrix operation expressed in matrix-vector form as $Y=SH$, where: S is a matrix of partial training bursts for each channel, each training burst segmented into N pieces in the time domain; Y is a vector of a received signal sequence; and H is a concatenation of different channel impulse response vectors.

25 10.   A method of estimating a channel impulse response comprising:-
         (a)   receiving a signal including a training sequence of predetermined training symbols,
         (b)   passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission
30   over a channel,
         (c)   demodulating at least one of the training symbols,
         (d)   calculating the non quadratic, non linear error metric of the demodulated training symbol against a locally stored record of its known correct value,
35   (e)   adjusting the estimated channel impulse response to substantially minimise the mean square error, and

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(f)   feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations

whereby the channel estimates used in the channel equaliser are iteratively refined.

11.   A method according to claim 10, wherein the training symbol error metric is calculated as mean square error.

12.   A method according to claims 10 or 11, wherein the received signal is a CDMA signal and wherein the training sequence is despread using at least one different spreading factor to recover training sequence bits or sub-bits.

13.   A method according to any of claims 10-12, wherein the channel equaliser is a Wiener (MMSE) equaliser.

14.   A method according to any of claims 10-13, wherein a first estimate of the channel response is generated by:-

performing transform operations on both a replica of the transmitted training sequnce s_n and a received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins;

performing   point-by-point   operations   between   corresponding   signal sequence frequency bins and training sequence frequency bins; and

concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the first estimate of the channel impulse response for the channel.

15.   A method according to any of claims 10-14, wherein the estimated impulse response is further refined by obtaining a long-term averaged power delay profile for the channel, setting a predetermined variance threshold, estimating a channel impulse response having a plurality of taps, and removing taps from the channel impulse response estimate equivalent to

56

those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

16.   A method of estimating a channel impulse response comprising:-

(a)   obtaining a long-term averaged power delay profile for the channel,

(b)   setting a predetermined variance threshold,

(c)   estimating a channel impulse response having a plurality of taps, and

(d)   removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

17.   A method according to claim 16, wherein the channel impulse response estimate is combined with its accuracy as predicted point-by-point from the combined power delay profile and interference level, whereby a modified short term channel estimate is produced in which points likely to have poor accuracy are deweighted and those likely to have good accuracy are retained such that final equalise performance is optimised.

18.   A method according to claims 16 or 17, wherein the channel impulse response having a plurality of taps is estimated by performing transform operations on both a replica of a signal sequence s_n and a received training sequence y_n received by the communication device in at least one burst, the received training sequence y_n being the signal sequence as received through a channel, the transform operations arranged to generate a multiplicity of signal sequence frequency bins and a multiplicity of training sequence frequency bins;

performing point-by-point operations between corresponding signal sequence frequency bins and training sequence frequency bins; and

concatenating the point-by-point operations associated with the channel to provide a composite frequency response for the channel, the composite frequency response allowing, in the time domain, generation of the channel impulse response for the channel.

19.   A method according to any of claims 16-18, wherein the channel impulse response estimates are refined by receiving a signal including a training sequence of predetermined training symbols, passing the received signal

through the channel equaliser to substantially remove distortion of the signal caused by transmission over a channel, demodulating at least one of the training symbols, calculating the mean square error of the demodulated training symbol against a locally stored record of its known correct value,

5   adjusting the estimated channel impulse response to substantially minimise the mean square error, and feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations whereby the channel estimates used in the channel equaliser are iteratively refined.

10

20.   A communications receiver arranged to receive signals from a plurality of sources and arranged to estimate a channel impulse response for the channel between the receiver and each source by:-

(a)   receiving a signal including a training sequence of predetermined

15   training symbols,

(b)   passing the received signal through a channel equaliser to substantially remove distortion of the signal caused by transmission over a channel,

(c)   demodulating at least one of the training symbols,

20   (d)   calculating a non quadratic, non linear error metric of the demodulated training symbol against a locally stored record of its known correct value,

(e)   adjusting the estimated channel impulse response to substantially minimise the mean square error, and

25   (f)   feeding the adjusted estimated channel impulse response back to the channel equaliser for use in subsequent equalisation operations

whereby the channel estimates used in the channel equaliser are iteratively refined.

30   21.   A receiver according to claim 20 including discarding channel estimate taps which have poor accuracy as predicted from the channel power delay profile.

22.   A receiver according to claim 20 or 21, wherein the training symbol error metric is calculated as a mean square error.

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23.   A communications receiver arranged to receive signals from a plurality of sources and arranged to estimate a channel impulse response for the channel between the receiver and each source by:-

(a)   obtaining a long-term averaged power delay profile for the channel,

5   (b)   setting a predetermined variance threshold,

(c)   estimating a channel impulse response having a plurality of taps, and

(d)   removing taps from the channel impulse response estimate equivalent to those which in the long-term power delay profile, have a variance below the predetermined variance threshold.

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24.   A receiver according to claim 23 including iterative adjustment of the channel estimates to minimise the mean square error on at least one training symbol.

1/8
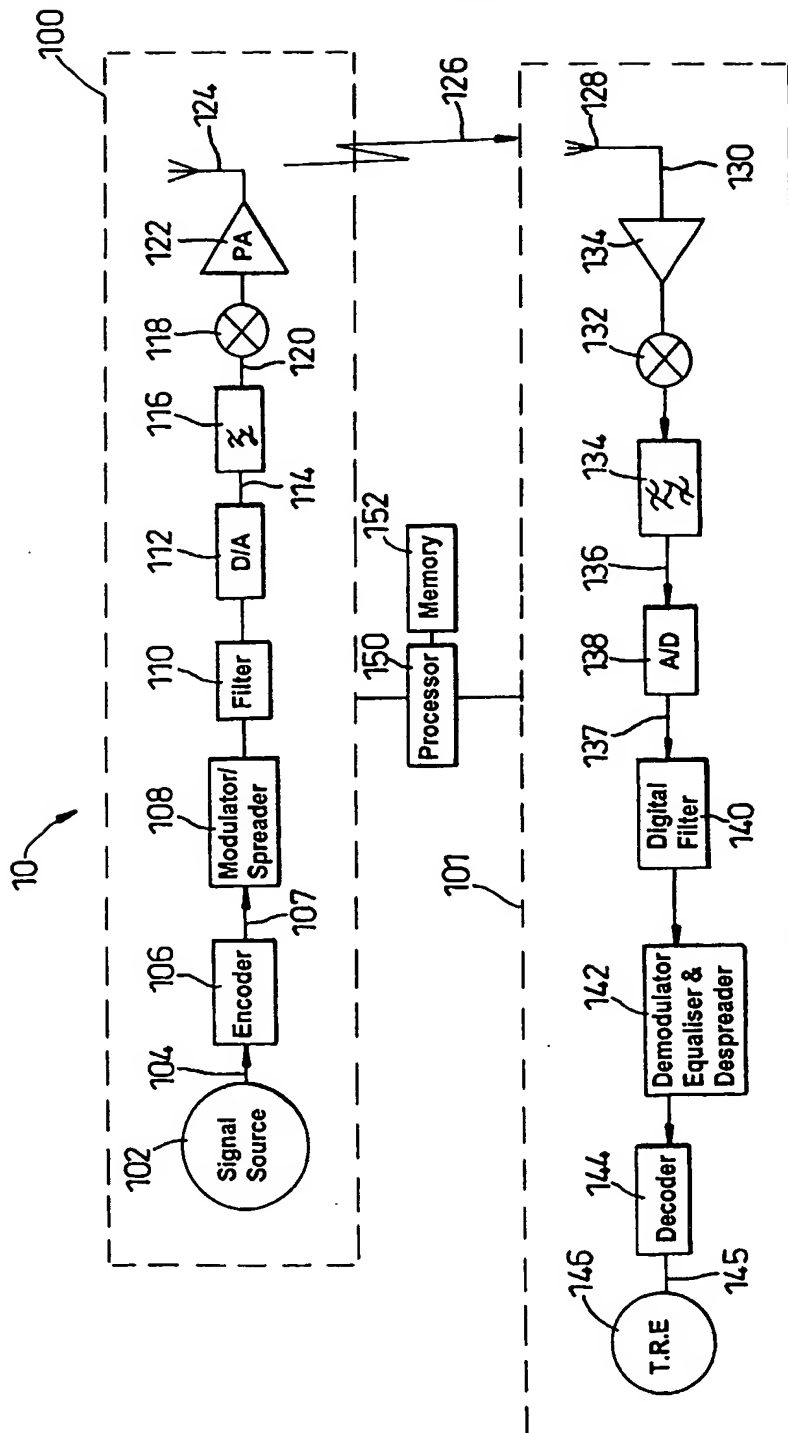
Fig. 1
Prior Art

2/8
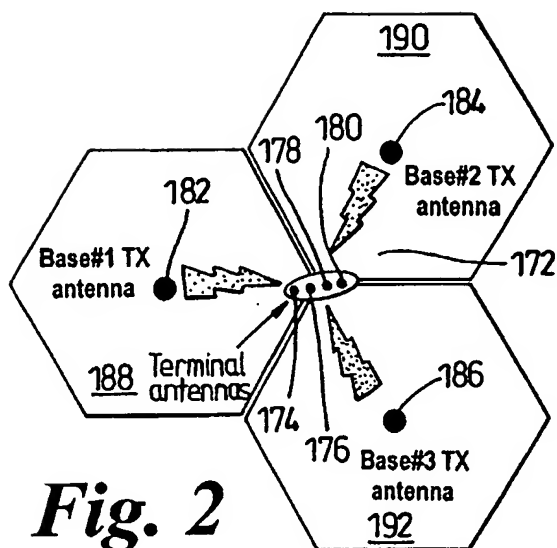
Fig. 2
Prior Art

Fig. 4

3/8
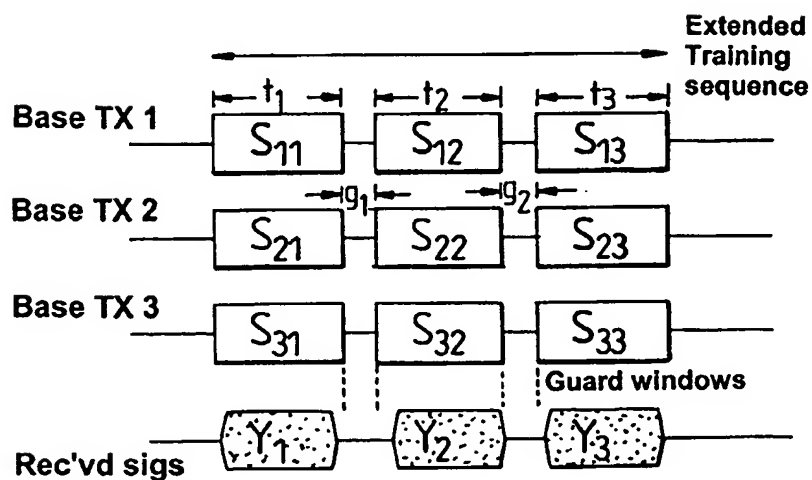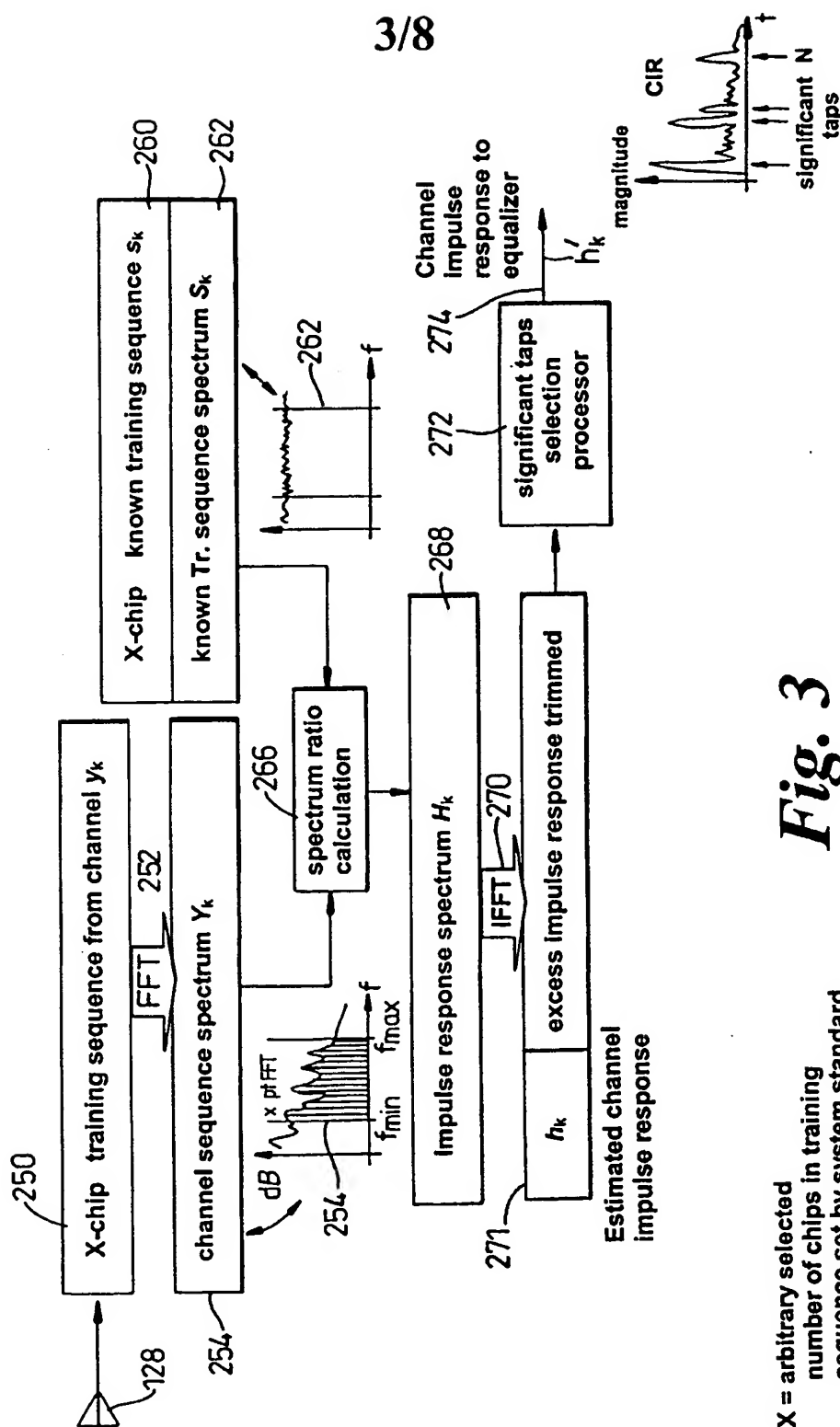
Fig. 3

X = arbitrary selected
number of chips in training
sequence set by system standard

<u>500</u>

Burst
#1

Burst
#2

$S_1(\omega)$ 502 $H_1(\omega)$ 508   512 $S_1^*(\omega)$ 514

BS1

$S_2(\omega)$ 504 $H_2(\omega)$ 510   513   Noise $N_1(\omega)$ $S_2^*(\omega)$ 516

BS2

506

520

$S_3(\omega)$ $H_1(\omega)$ 508   512 $S_3^*(\omega)$ 528

BS1

$S_4(\omega)$ 522 $H_2(\omega)$ 510   Noise $N_2(\omega)$ $S_4^*(\omega)$ 530

BS2

524

526

Frequency
domain
decoupler
518

$\hat{H}_1(\omega)$ 532

$\hat{H}_2(\omega)$ 534

Time

Fig. 5

602   604

burst 1   burst 2

Extended
Training
sequence

Base TX 1 $\{s_{11}\}$ $\{s_{21}\}$

Base TX 2 $\{s_{12}\}$ $\{s_{22}\}$

**Guard windows
or cyclic prefixes**

Rec'vd sigs $\{y_1\}$ $\{y_2\}$

Fig. 6

5/8

Fig. 7

6/8

Fig. 8

Fig. 9

Fig. 10

7/8

Fig. 11

Fig. 12

Fig. 13

8/8

Fig. 14

Fig. 15